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AIRCRAFT FIBER-OPTIC INTERCONNECT SYSTEMS PROJECT.(U)

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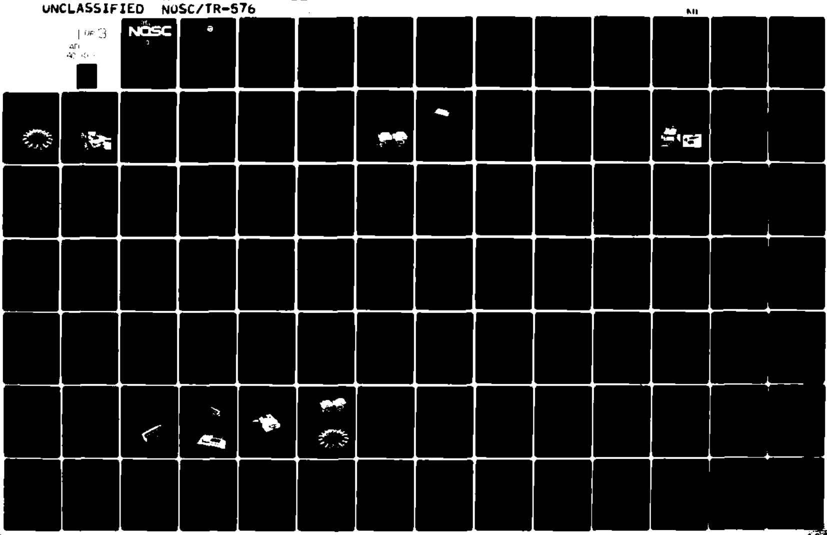

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AIRCRAFT FIBER-OPTIC INTERCONNECT SYSTEMS PROJECT.

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R.D. Harder/IBM Federal Systems Division

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NAVAL OCEAN SYSTEMS CENTER
SAN DIEGO, CALIFORNIA 92152

NAVAL OCEAN SYSTEMS CENTER, SAN DIEGO, CA 92152

AN ACTIVITY OF THE NAVAL MATERIAL COMMAND

SL GUILLE, CAPT, USN
Commander

HL BLOOD
Technical Director

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This report summarizes the development and fabrication of a fiber-optic interconnect intended for use in a MIL-STD-1553 multiplex system.

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INTRODUCTION

The Avioptics Program is an engineering and application development program whose central objective is to support the preparation of fiber-optic transmission technology for use in naval aircraft. The program was organized and managed by the Air Surveillance Systems Project Office (Code 7309) of the Naval Ocean Systems Center (NOSC). The Avioptics Program Plan is presented in detail in NOSC Technical Document 197.[1]

The applications of principal interest in the Avioptics Program are the transmission interconnects of aircraft multiplex systems. Included within the scope of this area are digital time division multiplex (TDM) systems, video multiplex and distribution systems, and such equipment ensembles as armaments, flight control, and voice communications.

A significant return on the investment in this program is expected to be realized in the reduction of life-cycle costs associated with aircraft electronic systems.[2] Factors such as reduction of shielding, filtering, and transceiver design requirements are anticipated together with enhanced resistance to electromagnetic disruption of system transmission. Fiber-optic transmission should raise the quality of system performance in severe mission environments above that achievable in conventionally wired systems under the same conditions. Fiber-optic technology also offers potentially wider transmission bandwidths than are possible with wire technology, a situation which could result in an expanded multiplexing capacity for a system of high-speed or high-quality signal transmission between points.

PROJECT DESCRIPTION

One of the key segments of the Avioptics Program is the Aircraft Fiber-Optic Interconnect Project which was performed by NOSC personnel and by personnel of IBM Federal Systems Division, Owego, NY, under contract N00123-77-C-0747. NOSC had responsibility for technical and administrative management of this contract.

This project was concerned with the interconnection of both low- and high-speed digital multiplex systems with fiber-optic transmission links. The project consists of two phases of effort: one is devoted to the low-speed digital application and the other to the high-speed digital application. Each phase consisted of an analysis and a supporting subphase of hardware design, fabrication, and testing. The low-speed digital system of interest is MIL-STD-1553.* The Phase II high-speed digital system simulated the projected data transfer requirements and operational features of a future, high performance avionic system such as VSTOL AEW or ASW. A fiber-optic interconnect was prepared, evaluated, and fabricated to serve the system's transfer requirements. A Draft Military Standard for a 50 MHz Free Access Multiplex Data Bus was prepared by IBM under this contract and is included in this report as Appendix C.

---

[1]NOSC TD 197, Avioptics Program Plan, by WJ Tinston, Jr, 25 September 1978

[2]NELC TR 1982, A-7 ALOFT Life-Cycle Costs and Measures of Effectiveness Models, by RA Greenwell, March 1976

Section 1

CONTRACT SUMMARY

Section 1

CONTRACT SUMMARY

IBM demonstrated fiber-optic data bus technology in an actual MIL-STD-1553 avionic system.

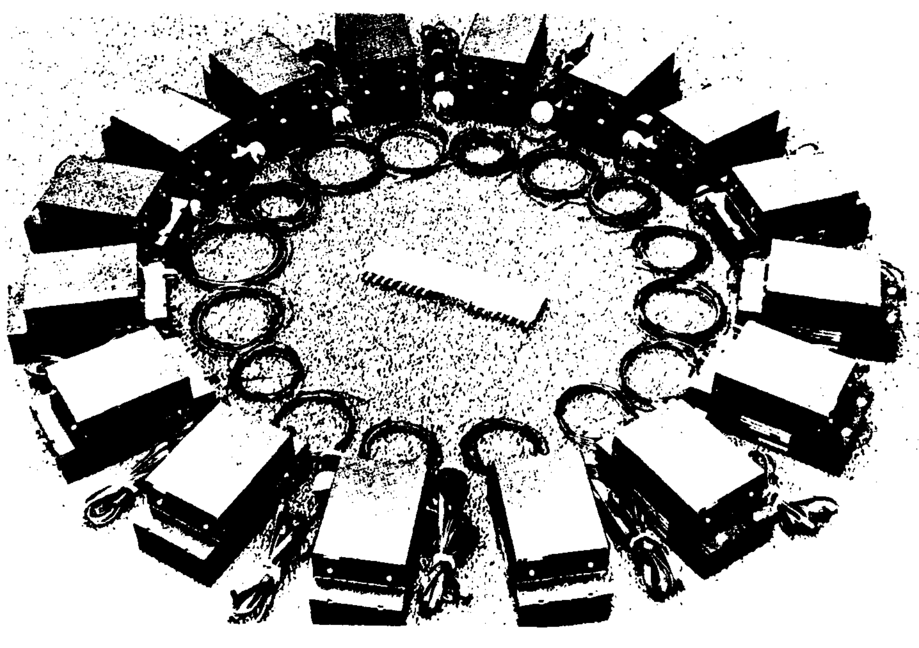
Key Accomplishments of Tasks I and Ia

Task I

1. Determined set of components to be used for the fiber-optic 1553B data bus.
2. Established optical modulation waveform (pulse position modulation, PPM).

Task Ia

1. Designed receiver, transmitter, encoder, and decoder.
2. Performed circuit packaging, fabrication, and test of the electro-optics.
3. Demonstrated equipment operating in a system in place of a conventional 1553 interface (error rate $< 1 \times 10^{-12}$).



Fiber-Optic 1553B Data Bus Electro-Optics

CONTRACT SUMMARY

IBM demonstrated future fiber-optic technology in a 50-MHz Manchester free-access-protocol data bus.

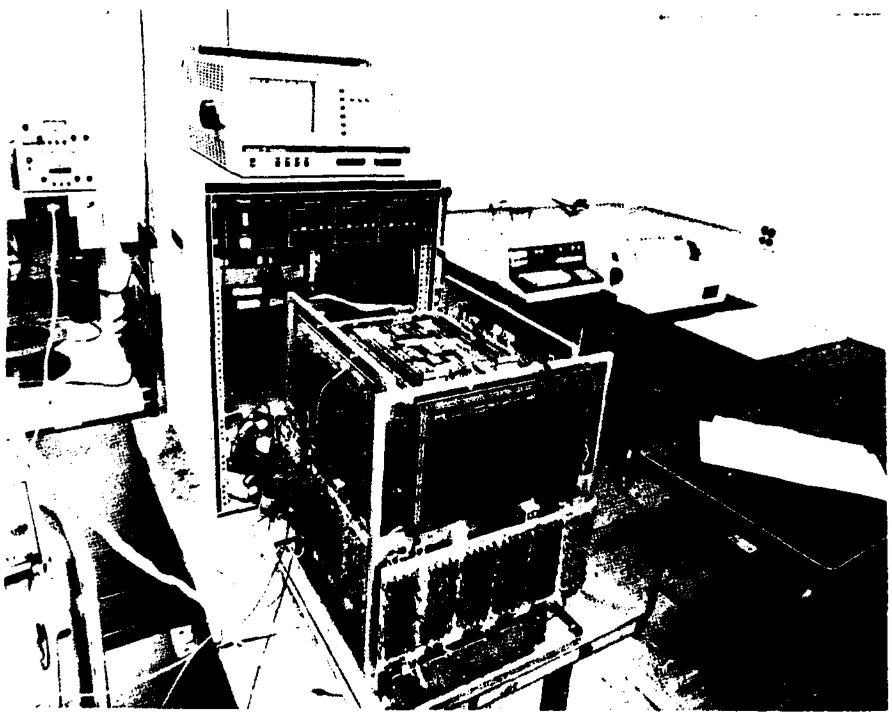
Key Accomplishments of Tasks II and IIa

Task II

1. Determined the future data transfer requirements for airborne early warning (AEW) and antisubmarine warfare (ASW) missions to be on the order of 12 megabits (Mb).
2. Determined 50 Mb is required to establish a Manchester-encoded, free-access-protocol, serial data bus with a 12-Mb information rate.

Task IIa

1. Designed and developed the electro-optics to function in a 100–m, 16–port, 50–Mb Manchester bus.
2. Designed and developed a demonstration system capable of exercising the electro-optics at full capacity.
3. Performed operational bus tests and demonstrated an error rate lower than MIL-STD-1553B at a bus information rate projected for the 1985-1990 time frame.



System Demonstration Equipment–50–Mb, Manchester, 16–Port Bus

CONTRACT OVERVIEW

Studies and demonstrations indicate that present and future applications of fiber optics will benefit naval avionic systems.

The Aircraft Fiber-Optic Interconnect Systems contract investigated two technological timeframes:

- Present – Technology as it is today
- Future – Projection to 1985-1990

The potential benefits of fiber optics are reiterated in the table opposite.

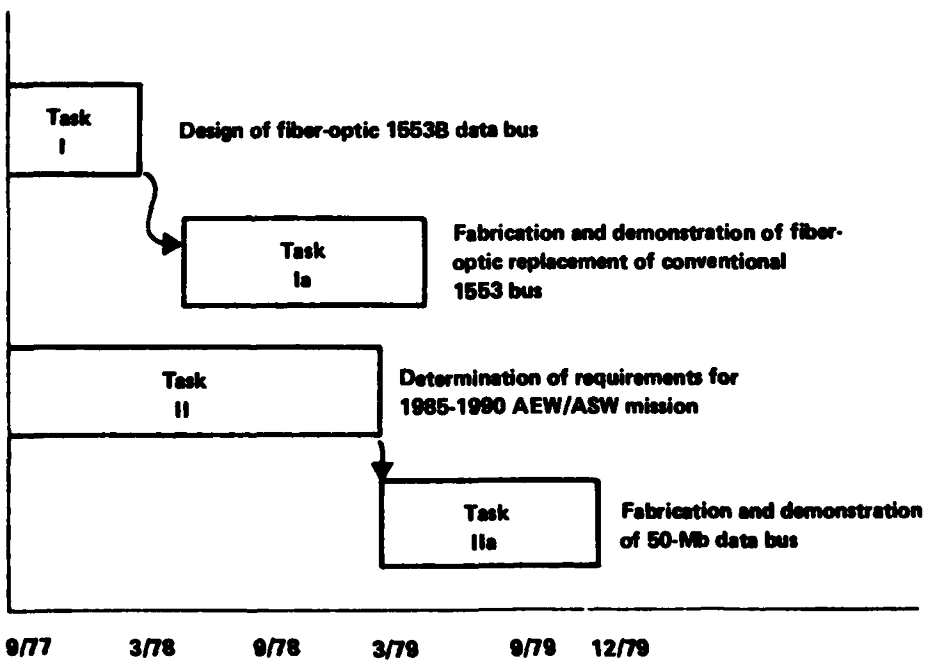
The contract had two basic tasks:

- **Task 1** - A study to determine the best available technologies to implement a MIL-STD-1553B-compatible, 1–Mb, fiber-optic data bus.

Task Ia was to build a laboratory breadboard of the data bus resulting from the IBM Task I study in a configuration approved by the Navy. The original contract assumed only a few representative terminals would be built, but a later addition funded the building of a complete 16-port system with all terminals.

- **Task II** - A relatively open-ended study, in that it required IBM (1) to look ahead 5 to 10 years and project what a "typical" Navy avionic system would look like at that time, and (2) to recommend in what areas fiber-optic technology could be applied to implement system and subsystem interconnections.

Task IIa was the building of a laboratory system to demonstrate these interconnects. IBM recommended, and the Navy approved, substituting certain technologies in order to make the demonstration feasible in the 1979 time frame.



Contract Time and Flow Diagram

Benefits for Naval Avionic Systems

Feature	Advantage	Benefit
Bandwidth	Not only are attenuations relatively low, but they are independent of frequency up to hundreds of MHz for lengths typical in aircraft.	Applicable to virtually all avionic data transmission needs.
Electro-magnetic Compatibility	Photons, unlike electrons, are not susceptible to, nor do they generate, electromagnetic fields.	Solves EMC problems. New systems can be introduced to an aircraft without risk of interfering with existing systems or failing to operate in the presence of existing systems.
Nuclear Radiation Hardness	Electromagnetic pulse (EMP) has no effect on the transmission medium. The level of neutron and gamma flux sensitivity is extremely high for plastic-clad, fused-silica fibers.	No premium expense in dealing with nuclear radiation hardness requirements.
Communication Security	Because of the absence of radiated fields, the system is secure, in that fiber must be penetrated in an attempt to covertly monitor the information; a condition that can be noticed at the terminals if designed to do so.	Security can be achieved at little additional expense.
Safety	Broken fibers are not dangerous in explosive or flammable atmospheres (no sparks)	The medium can be freely routed through fuel compartments, oxygen areas, or any other atmosphere dangerous for electrical wiring.
Weight	At frequencies above 10 MHz, serial data buses are extremely difficult, if not impossible, to build with wire. It follows, then, that potential weight savings exist when a single-fiber cable with no metal shields can replace several parallel wires having shields. In addition, long runs of fiber cable weigh much less than the equivalent in coaxial cable.	The weight savings on a typical advanced avionic system can be very significant.

SUMMARY OF TASK I RESULTS

Task I established the design of a MIL-STD-1553B fiber-optic data bus.

Task I addressed those elements of the data bus that had to be designed, or for which components had to be chosen:

- Sources (primarily LEDs)
- Fiber-optic cables
- Cable connectors
- Optical couplers
- Detectors (primarily PIN diodes)
- Optical modulation waveform
- 1553-compatible encoding and decoding
- Transmitter and receiver designs

Element Selection

Tradeoffs were based on achieving the optimum system, which is not necessarily synonomous with optimizing each element. IBM imposed the requirement for each element to meet, or show evidence of being able to meet, MIL-E-54 0D, Class II, environments.

Because of the maturity of the devices and the designs using them, LEDs and PIN photodiodes were selected almost immediately. The initial optical losses estimated for the data bus virtually guaranteed an adequate design. The only available LEDs capable of meeting the military hermeticity specifications were large-area emitters. Hermetic coupling to single-fiber cables via pigtails could not be demonstrated. Without pigtails, unacceptable input coupling losses exist between the LED and the cable. Consequently, fiber bundles of 1 mm to 1.2 mm diameter had to be used. This, in turn, determined the physical size of the couplers, connectors, and detector chips.

Modulation Waveform

Detailed theoretical studies of the optical modulation scheme giving the best bit error rate performance were made, and it was concluded that a pulse position modulation (PPM), encoded from the logic-level Manchester of MIL-STD-1553, would be optimal.

Basically, the scheme required a 50-ns, high-intensity, optical pulse to be generated for each 500 ns of up, or "on", time of the normal Manchester-encoded signal. On the average, the duty cycle of this modulation is one-tenth the duty cycle of the full-width Manchester. It was possible, therefore, to increase the amplitude of the optical pulse by ten times and still maintain relatively the same heating effect in the LED junction and not degrade LED reliability. This effect has been studied theoretically by Dr. Baird of Spectronics and will be discussed in later sections.

The final result of the task I studies is shown in the table opposite. Section 2 gives detailed descriptions and analyses of possible configurations leading to this recommended data bus.

Recommended Data Bus

Source	Spectronics 2231 LED, or equivalent
Fiber	PCS 215-μm core; 10-40 dB/km
Cable	Reinforced, heavy-duty, 19-fiber bundle
Connector	Crimped, hex packed, keyed $\pm 5^{\circ}$
Optical Waveform	"PPM Manchester," 50-ns pulse
Receiver	Transimpedance-amplifier preamplifier, with limiting-differential-amplifier postamplifier, with tailored bandpass filter
Transmitter	LED driven by 1-A, 50-ns fast-rise time pulse
Coupler	16-port passive transmissive

Note: The actual system demonstrated required a change of fiber-optic cables.

SUMMARY OF TASK Ia RESULTS

Task Ia demonstrated a 16-port, fiber-optic, MIL-STD-1553B data bus

For Task Ia, IBM fabricated, tested, and delivered a 16-port, optical, data-transmission system. Each terminal contained its own power supplies, a transmitter a receiver, and logic that provided an interface for the proposed MIL-STD-1553B fiber-optic data bus.

Interface

The fiber-optic interface is illustrated, along with the waveforms proposed for use between the Fiber-Optics Transmit/Receive Unit (FOTRU) and a bus interface unit (BIU).

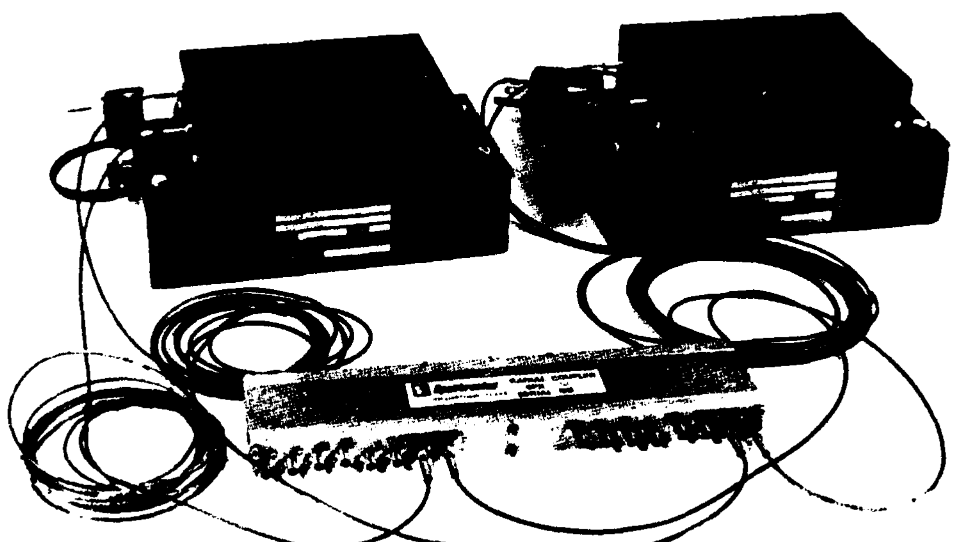
The BIU requires a positive Manchester logic-level (TTL) waveform (M) as input, and makes available a negative Manchester logic-level waveform (P). along with a second negative waveform ($\overline{N}$), used to generate the three-state waveforms if required. The TTL levels figure shows the output waveforms available at the interface of the IBM terminals and the input waveform required for operation.

IBM's FOTRU supplies two signals in addition to that supplied by the proposed FOTRU; that is, a plus threshold and a minus threshold. In some existing equipment, they are also required to be compatible with previous 1553 A/B wire standards.

Packaging

IBM decided to package the transmitter and receiver in hybrid integrated-circuit form, and the encode/decode logic on cards using DIPs. Those were put together with small, commercially available power supplies into a self-contained terminal that uses power from a 115-V, 60-Hz source.

A 16-port transmissive coupler was procured, along with several of the optical cables utilized in the final demonstrations. The photograph shows the ensemble of terminals, cable, and coupler used in the system data bus demonstration.



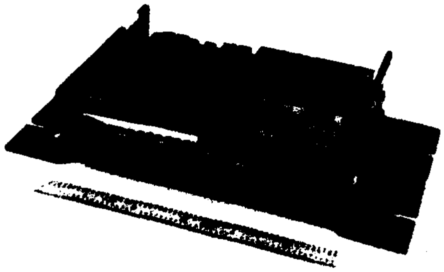
FOTRU Couplers and Cables

Demonstration

For the final demonstration, IBM had in place a Systems Integration Facility (SIF) Laboratory, which was a mockup of an avionic navigation, flight control, and weapon delivery system. The SIF contained real and simulated (IBM System/370 computer) hardware and had been interconnected via a double redundant 1553 wire bus. One of the redundant wire buses was replaced with the Task Ia fiber-optic hardware, and the two were used alternately, with equally successful results.

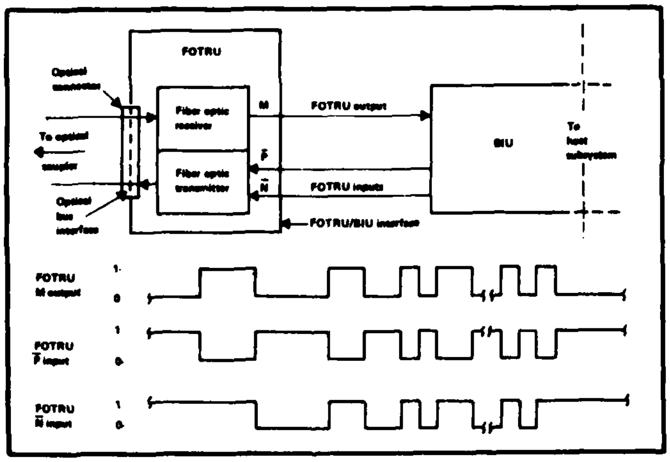
Bit Error Rate

In addition to the system demonstration, a bit error rate test was run using the fiber optics. An extrapolated error rate of 1×10^{-12} was achieved. All system and hardware tests are detailed in Section 3.

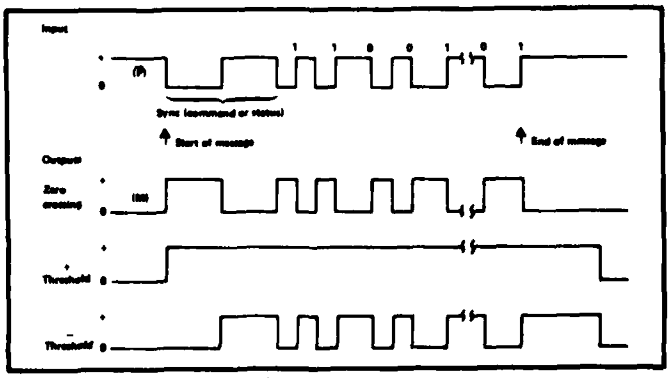


Left-hand side of photo shows logic packaging; the right-hand side, the hybrids.

Standard Single-Frame EECO Package



Fiber-Optic Interface and Associated Waveforms



Manchester TTL Level Waveforms

SUMMARY OF TASK II RESULTS

Task II of the fiber-optic bus study determined the data bus requirements for future naval avionic systems.

IBM analyzed the intended mission scenarios of the Navy's future aircraft and defined the mission functions and total weapon system architecture, to derive operational requirements for the interconnection system.

Configuration Determination

From the analysis and definition followed the avionic system, the information-handling architecture, the data flow, and, finally, the interconnection system operational requirements.

To develop the "worst-case," or most complex, requirements, two types of missions were studied: airborne early warning (AEW) and antisubmarine warfare (ASW).

The block diagram of a "typical" future system illustrates the mix of future processors. That system was configured as a "distributed system," relative to its command and control requirements, whose distribution is implemented by the bus at the top of the figure. The heavy lines represent bulk data transfer requirements that take the form of dedicated links and storage pool buses. The avionics scan bus shown at the bottom of the figure does the normal flight control and stores management intercommunicating.

Analysis showed the information rate requirements for the distributed command/control bus to be on the order of 12 Mb/s. The bulk transfer bus will require rates in the 300- to 400-Mb/s range. The avionics scan bus is not seen as requiring anything greater than the current 1553B rates and protocol.

Protocol

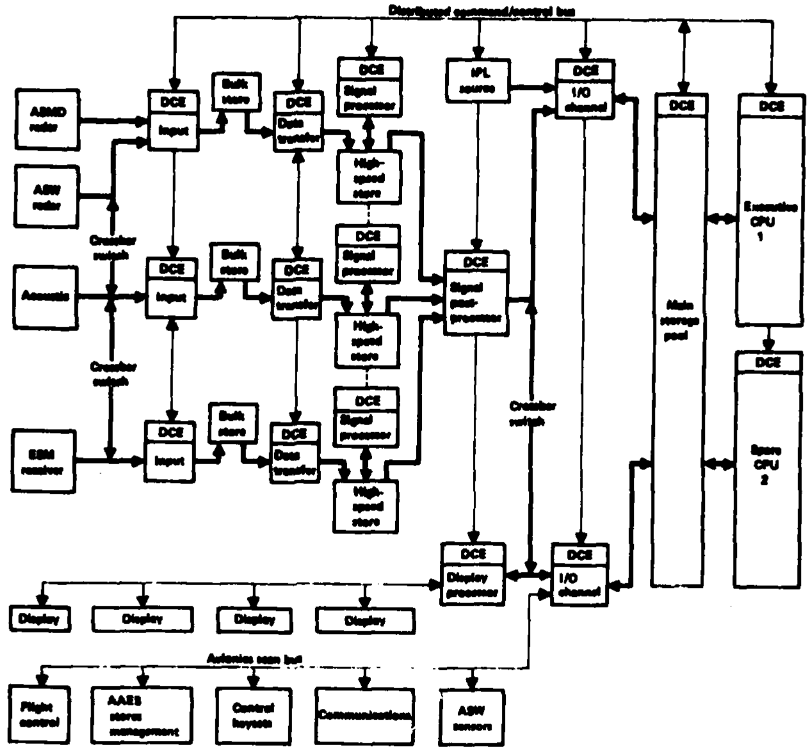
Because of the distributed nature of the command control bus, a versatile protocol is required. IBM chose to recommend a free-access protocol, with all terminals contending on a priority basis at any given time. This choice was considered best overall, due to its flexibility. It was recognized, however, that this kind of protocol is less efficient than others and requires a greater raw bit rate capability in the bus hardware, in order to account for the protocol and for delays between remotely located terminals.

The information-rate figure demonstrates the real information rate of a bus capable of raw bit transfers of 50 Mb/s, Manchester encoded. The real information considered is only the 16 bits of data in a 20-bit word, where 3 bit times are taken for word sync and 1 bit is for parity. Time not contributing to information transfer is all the protocol bit time, the contention time, the delay time, sync, and parity bit times. The curve is plotted for one of 16 terminals on a 16-port bus, where the terminal is the one with the median priority. For low average message lengths, the real information rate varies considerably with bus length, because delay time is a significant portion of the total time at the data rates considered here.

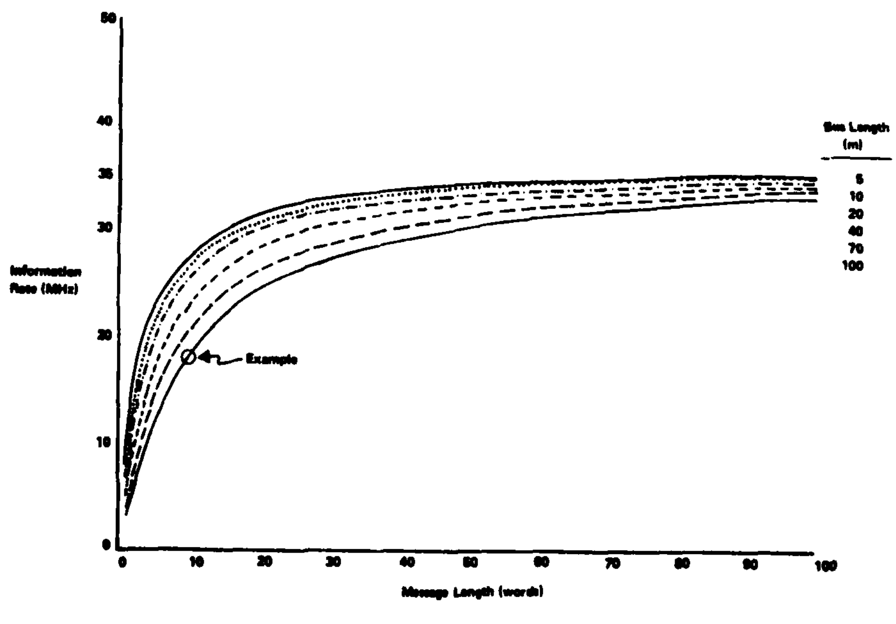
Conclusion

Having considered the alternatives, IBM recommended that Task IIa be devoted to the development of a 16-port fiber-optic data bus, with a raw bit rate capability of 50-MHz Manchester, a maximum terminal separation of 100 m, a word error rate of less than 1×10^{-7}, a bit error rate of less than 1×10^{-12}, and a free-access protocol. Higher data rate buses must be achieved using parallel buses, because it was felt that 50-MHz Manchester was stretching, not only bus technology, but the logic technology required in the terminals.

Much greater detail of the system studies and analyses leading to these conclusions is given in Section 4.



"Typical" Future System



Real Information Rate

SUMMARY OF TASK IIa RESULTS

Task IIa demonstrated a 50-MHz, fiber-optic, distributed-control, free-access data bus having 16 ports. The block diagram shows the demonstration system.

Three identical message units were designed and built. The first two each represented one of 16 terminals on a 16-port bus; the third simulated the other 14. A message operating system, along with an IBM 5110 computer, ran the system in various configurations and collected and analyzed data to determine system operating characteristics.

The effort was subdivided into four major design tasks: (1) the optical bus (fiber-optic transceiver, optical coupler, and cable), (2) the encoding, decoding, and protocol logic (bus operating system and Manchester encoder/decoder), (3) the demonstration control and data-handling logic (message operating system and IBM 5110 computer), and (4) the software required to produce the test data and system operating characteristics.

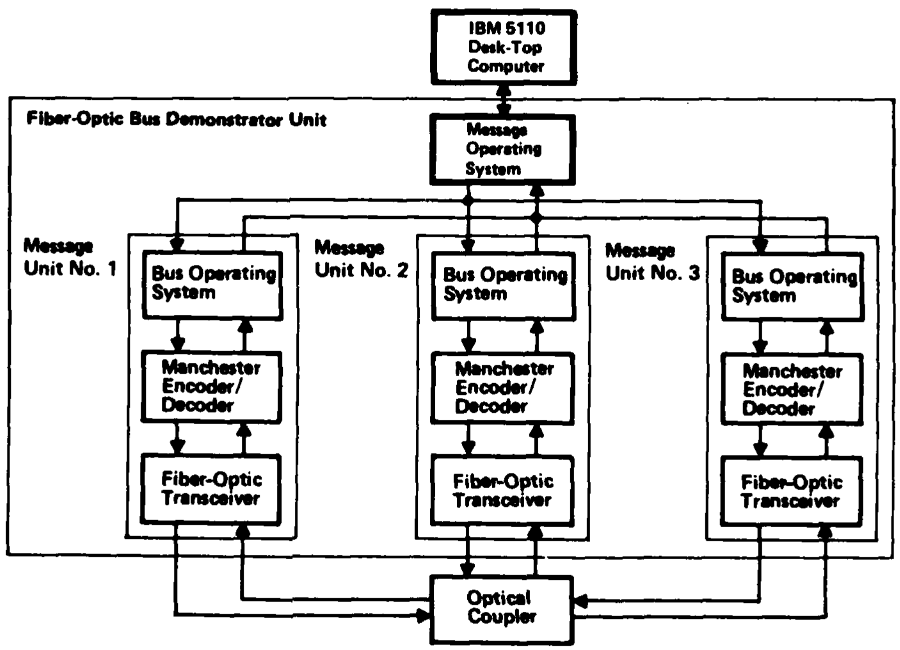
The system was designed, fabricated, and debugged; and then was successfully demonstrated on 11 December 1979.

The performance table summarizes the actual results achieved and compares them to IBM's goals. The actual equipment is shown in the photographs.

The components table lists the optical components orignally recommended and those finally procured. The digital logic portion of the equipment was implemented with standard MSI TTL and 10K series emitter-coupled logic.

Much greater detail of the design methodology, tradeoff analyses, testing procedures, and results can be found in Section 5.

IBM believes that the conclusions drawn from this effort enable the confident projection of a fully qualified MIL-STD fiber-optic, 50-Mb, 16-port, serial avionic data bus in the 1985-1990 time frame.



High-Speed Fiber-Optic Data Bus Demonstration System

High-Speed Bus Performance

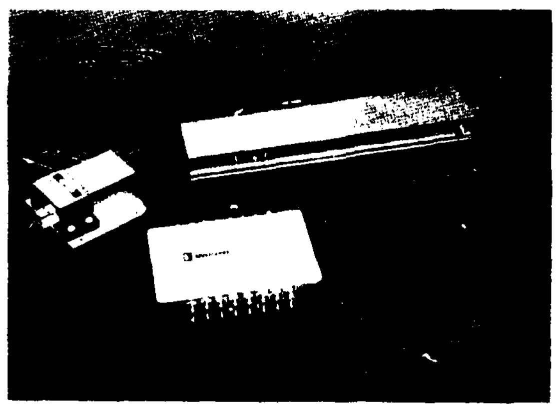
Item	IBM Goals	Achieved (1979)	Projected (1985-1990)
Bit rate (serial)	50 MHz	50 MHz	50 MHz (minimum)
Number of terminals	16	16	16-32
Maximum terminal separation	100 m	> 100 m	> 100 m
Raw bit error rate	1×10^{-12}	1×10^{-12} 6.4×10^{-10} system	1×10^{-12} 1×10^{-12}
Incomplete message rate	1×10^{-7}	*1.4×10^{-8} per word	$< 1 \times 10^{-8}$
Timing	Asynchronous	Yes	Yes
Protocol	Free access	Yes	Yes

*Equivalent to 1.4×10^{-8} word error rate (1553B requires 1×10^{-7} word error rate)

Optical Components

Component	Recommended	Actual
Source	Pigtailed LED Rise time 2-3 ns Power out ~ 500 μW into 200-μm core 800-900 nm	GOI 1-3 laser transmitter < 1 ns 4.5 mW/V 830 nm
Detector	PIN photodiode Rise time ⩽ 1 ns Responsivity ⩽ 0.5 A/W	Spectronics PIN photodiode < 1 ns at 15 V 0.23 A/W
Fiber (cable)	Plastic clad, fused silica, step, single 200-μm core ⩽ 6 dB/km ⩾ 25 MHz-km	Glass clad (Gallite 3000) 204-μm core 50 dB/km 25 MHz-km
Coupler	16-port transmissive ⩽ 16-dB loss without connector Small structure Pigtails	Spectronics 16-port transmissive Type I, 19.2 dB. Type II 15.7 dB Type I Connectors Type II Pigtails
Connector	Single fiber ⩽ 1.5-dB loss	Single fiber ITT 1-dB loss

Terminal, Encode/Decode, Rx



Fiber-Optic Receiver, Transmitter, and Star Coupler

LESSONS LEARNED

The described contract had two main outcomes: (1) it has clearly demonstrated the maturity of fiber optics for interconnecting avionic equipment; (2) it has developed specific designs and architectural concepts for similar near-term and future applications.

Conclusions and Recommendations from Tasks I and Ia

- For an optical MIL-STD-1553B, 1-Mb/s data bus, there exists today a compatible set of components that can be qualified to MIL-E-5400 and can be used to implement the design.
- Until the hermiticity problem of pigtailed LEDs and photodiodes is solved, bundled fiber cables should, and can, be used.
- The form and implementation of the PPM optical waveform described in this report can give a 4- to 5-dB margin over a full-width Manchester modulation.
- If a solution to the plastic-clad, silica-fiber termination problem can be found, the losses in a bundle-fiber connector can be reduced to less than 2 dB (see Section 3.2).
- More development should be focused on improving the design of bundle-fiber couplers if pigtailed components cannot be used for military environments.

Conclusions and Recommendations from Tasks II and IIa

- Without use of an avalanche photodiode (APD) detector at the front end of the receiver, currently available LEDs are neither fast enough, nor do they emit enough energy, for a 50-Mb/s, Manchester-encoded data bus. Even with the APD, the operation would be marginal.
- A laser transmitter is required, but it should be one designed for digital signals. That will require optical feedback to change the average light output from the threshold level with no signal to some average level in the presence of a Manchester-encoded digital signal, or to some maximum level if NRZ encoding is employed.
- Improvements can be made in the receiver design to minimize the waveform transition timing ambiguity. Two-sided differential amplifiers should be used, and the comparators should be driven from a collector that is turning on.
- Cable terminations could still be a problem, based on the failure rate of those purchased for this task. This may be an isolated instance, but it should be carefully watched.
- The pigtailed 16-port coupler was superior to the coupler with connectors, as was expected. This is because the internal construction is much simpler. Where an installation will remain fixed for long times, it should be the obvious choice.
- The use of wirewrapped ECL 10K logic seems to have caused some problems with encoder/decoder design. In particular, those areas operating at 100 MHz were very susceptible to the logic net construction and terminating resistor voltage. We would expect this implementation of the encoder/decoder to be very difficult to reproduce in quantity, because each unit must be individually adjusted. Therefore, we recommend that future encoder/decoder units use ECL 100K flatpacks mounted on printed circuit (PC) boards, which would result in an easily reproducible unit that operates reliably.

  Note that changes are much more difficult on an ECL PC board, because softwires must be semirigid coaxial cable. Therefore, more effort should be put into the design phase than for a wirewrapped ECL design.
- The current decoder produces more bit errors than desired, because its clock is asynchronous with the data coming in. That occasionally causes insufficient setup time in certain critical flip-flops, causing bit errors. Although certain design modifications in the decoder could reduce the frequency of the bit errors, the best solution for future serial bus systems is a new decoder design. The new design can obtain all of its timing from the received waveform rather than from a fixed-frequency oscillator.
- There is no doubt that, for contracts starting in the time frame of 1985-1990, a 50-Mb/s, 16-port, Manchester-encoded, serial, digital fiber-optic data bus can be built to meet the full military environments of MIL-E-5400X, Class II.

Section 2
TASK I DETAILS

This section details the point of departure, the study methodology, the tradeoffs made, the recommended system, and the test plan for deomonstrating the capability to instrument a MIL-STD-1553B data bus with fiber-optic technology.

2.1 INITIAL ASSUMPTIONS

IBM and the Navy agreed on a groundrule to require the selection of components and technology already available which could meet the pertinent specifications and test parameters for a MIL-E-5400D, Class II environment. Althrough the contract did not fund or require extensive qualification of parts or subsystems, the equipment was designed as if it would be subjected to such tests. A further requirement or groundrule was that the interface between the fiber-optics and the 1553B protocol logic would (1) permit interchanging a transformer-coupled, twisted, shielded pair bus with a fiber-optic bus, and (2) be transparent to the user system.

2.2 TASK I METHODOLOGY

Figure 2-1 is a flow diagram of the approach to the Task I study.

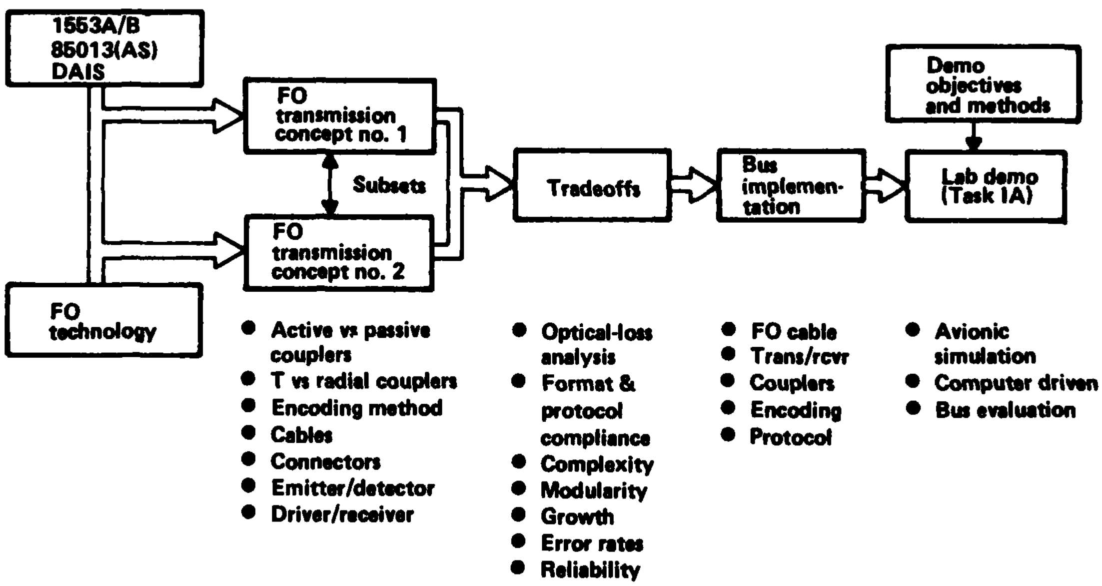


Figure 2-1. Task I Methodology

Based on IBM's 1553A and DAIS experience, the 85013(AS) Multiplexing and other specifications, and the latest fiber-optic technology, IBM explored at least two transmission concepts using various subsets of possible hardware. Hardware variables included (1) active or passive bus coupling, (2) "T" or radial passive couplers, (3) single or bundle fiber cables, (4) emitters, (5) detectors, and (6) various transmitter and receiver designs. System variables included encoding and decoding data, system logic interfaces, and optical modulation schemes.

The two transmission concepts underwent tradeoffs using criteria such as optical loss, 1553 format and protocol compliance, complexity, modularity, growth capability, bus error rates, reliability, etc. These tradeoffs and analyses, detailed next, led to the recommended system implementation and the test/demonstration plan presented to the Navy.

2.3 COMPONENT AND BUS SYSTEM TRADEOFFS

A survey of fiber-optic cables, light sources, photo detectors, optical connectors, multiport couplers, and Rx/Tx modules determined the new technology of available components as of January 1978. The survey included contacts made via telephone, trade shows, and visits to manufacturers. Over 125 companies were contacted, including every known source in the U.S. and many foreign companies.

More companies are now active in the marketplace, but no significant changes have occurred in the past 2 years. Most of the new manufacturers are looking at long-line, single-fiber, point-to-point telecommunications markets. The array of available components has no standard guidelines, and each manufacturer views its component as a stand-alone item, not a system component. The result is "apples and oranges" combinations for a system user. The system designers find that widely advertised specifications and claims of the manufacturers for individual components do not integrate at the system level.

2.3.1 FIBER-OPTIC CABLE

The fiber-optic cable is the common point of a fiber-optic system. All the components of the system have a relationship with the fiber optics characteristics and, therefore, the parameters of the fiber-optic cable must reflect the best system solution or compromise. Thus, the particular system sought may cause the desired specifications of the fiber-optic cable to vary with the application. The application here is a multiport serial data bus applicable to military avionic systems. A more specific description of the system's general characteristics is a serial data bus patterned after MIL-STD-1553, using fiber-optic transmission media.

Fibers are generally fabricated coaxially of materials that produce an index variation between the inner and outer materials, with the lowest index occuring in the outer material. This structure forms a dielectric waveguide that guides light within the fiber. The structure is generally made using glass-on-glass, plastic-on-silica, or plastic-on-plastic material structures. Plastic-on-plastic cannot meet the military aircraft environment (MIL-E-5400) because of adverse temperature and humidity effects on the plastic core, excluding this type from this application.

The properties of the dielectric waveguide are a function of the materials used and the waveguide's geometry. The two general properties derived from these parameters are loss per unit length and pulse dispersion per unit length. Both are somewhat correlated because they are related to the number of modes allowed to propagate in the waveguide. In general, high-order mode propagation produces higher losses and pulse dispersion. The lowest loss and least pulse dispersion occur when only a single mode is allowed to propagate.

The numerical aperture (NA) of a fiber is a measure of the number of modes propagated. Fibers with high NAs (0.4 to 0.7) usually have high losses (100 to 1,000 dB/km); fibers with low NAs (0.1 to 0.2) have low losses (1 to 10 dB/km). Fibers having losses in the medium-loss category (10 to 100 dB/km) have NAs in the range of 0.2 to 0.4. Apparently, this medium-loss category of fibers with its moderate pulse dispersion characteristic would best fill the needs of a serial avionic data bus. This type of fiber was surveyed.

Table 2-1 is representative of materials being offered by fiber manufactures in the medium-loss category. These data represent the best offer of seven companies that supply fiber-optics for communications. The two general categories of glass-clad glass and plastic-clad silica (PCS) are represented. Note that glass-clad glass fibers have core diameters from 50 to 100 μm and PCS fibers have core diameters from 125 to 250 μm. Note also that fiber cables with multiple fibers are available from most of these companies but never with more than 61 fibers. Most companies produce bundles with 7 or 19 fibers.

Table 2-1. Survey of Fiber-Optic Cables - January 1978

Manufacturer and Products	Model	Core Diameter (μm)	Clad Diameter (μm)	NA	No. of Fibers
Galileo					
Glass-clad glass	Gallite 3000	89	110	0.48	1-7-19-37-61
Plastic-clad silica	Gallite "FAT"	204	245	0.35	1-7-19-37
Valtec					
Glass-clad glass	MS05	65	125	0.26	1-7-19-37
Plastic-clad silica	PC05	125	200	0.3	1-7-19-37
Plastic-clad silica	PC10	250	400	0.3	1-7-19
Quartz Products					
Glass-clad glass	NGS-SI-100	100	150	0.35	1
Plastic-clad silica	QSF-A-200	200	400	0.22	1
Dupont					
Plastic-clad silica	PFX-S-120R	200	600	0.4	1
Plastic-clad silica	PFX-S-220R	200	600	0.4	2
ITT					
Glass-clad glass	GS-02-8	50	125	0.25	1-6-7-19
Plastic-clad silica	PS-05-20	125	300	0.3	1-6-7-19
Times					
Glass-clad glass	SA10-90	90	125	0.16	1-3-6-10
Corning					
Glass-clad glass	Corguide "FAT"	90	125	0.3	1-6-7

2.3.2 LIGHT SOURCES

The array of sources available is related to the array of fibers on the market. All sources have certain commonality; they are typically GaAs devices or GaAlAs devices and emit at IR wavelengths from 750 to 950 nm. This is where commonality ends. Instead of centering on any specific type, Table 2-2 lists the entire spectrum of types to point out the various categories of sources that have evolved.

Table 2-2. Survey of Light Sources - January 1978

Company and Model	Type
Spectronics	
SPX-XXXX	Reflector-enchanced edge emitter
Laser Diode Labs	
IRE-140	16-mill surface emitter
IRE-150	9-mill edge emitter
IRE-160	8-mill etched well
IRE-152	4-mill edge emitter
IRE-103	Stripe geometry/LED
LCW-5	Strip geometry laser
Texas Instruments	
TIXL-XXX	Hemispherical domed emitters/reflector
RCA	
C-30119	Low-power double-het LED
C-30123	High-power double-het LED
C-30133	High-power double-het LED with fiber
C-30127	Low-power laser
C-30130	High-power laser
ITT	
801-E	Stripe LED
851-S	Double-het surface emitter (hole)
901-L	Stripe laser
Hitachi	
HLP series	16-mill hemispherical domed emitter

The Table 2-2 data represent categories of sources offered by companies supplying the communication industry. The list does not indicate the entire product line of the companies shown but simply the general category of device types offered.

Spectronics makes a variety of edge-emitting diodes that are optically enhanced by a reflector. No specific part number is referenced because the array of parts available from this company is a variation of chip size and reflector quality. Sources appropriate for use with high-loss and medium-loss fibers have been produced by this technique.

Laser Diode Laboratories (LDL) makes a variety of devices without external optical enhancement. This company has, however, the widest variety of source types, including the surface emitter, the edge emitter, the etched-well emitter, and the stripe-geometry cavity emitter. Only the stripe-geometry cavity devices have internal optical enhancement. The other sources, regardless of specific output power, have essentially Lambertain output radiation patterns, producing high input coupling losses, especially when used with medium- or low-loss fibers.

LDL's stripe-geometry sources used as LEDs or lasers produce low coupling losses if a single fiber is placed near the emitting surface. Two problems, however, have resulted: (1) the fiber core and the emitting surface of the diode are both very small and location and registration of the two is extremely difficult, and (2) the hermetic seal of the resultant pigtail has not yet been perfected.

Texas Instruments makes a variety of hemispherical domed emitters, some of which are optically enhanced with a reflector. The hemisphere allows more light to leave the chip, but the radiation pattern is nearly Lambertain. With this pattern, only those devices with external reflectors have low coupling losses. The domed emitter in a reflector is difficult to manufacture because of (1) the intermediate chip required to mount a hemispherical diode, and (2) the critical location of the chip and its mount in the reflector.

RCA recommends sources of the double hetero-junction cavity type only for communications. Their product line is merely a selection of quality and size of the same junction type. These sources have internal optical enhancement and can produce low coupling losses, but because of their small size, they have a rather severe alignment problem. These devices can be supplied with a fiber epoxied in place but not hermetically sealed.

ITT makes a series of cavity lasers. They supply a stripe laser or LED with or without a pigtail. They also have a cavity device with a hole etched on the cavity side, which is similar to an etched-well emitter and has a Lambertian radiation pattern. It also is supplied with or without a pigtail. None of the pigtail devices is hermetically sealed, but if they were, they would all have high coupling losses.

Hitachi makes a domed emitter with or without a pigtail. Its lack of optical enhancement produces high coupling losses. The package is also not hermetically sealed.

In summary, most manufacturers are addressing the single-fiber-source market, and many devices do not have optical enhancement. None of the pigtail fiber devices is hermetically sealed.

2.3.3 PHOTODETECTORS

The detectors available for optical communications systems are PN, PIN, and avalanche photodiodes. There are various enhancements such as coatings and guard rings that optimize certain device parameters, but there is no particular outstanding differences between devices. The silicon PIN photodiode was chosen for our application because of its simplicity of application, low noise, low capacitance, and high speed of response.

Table 2-3 is representative of material being offered by photodiode manufactures in the Silicon PIN photodiode category.

Table 2-3. Survey of Photodetectors - January 1978

Company and Model	Company and Model
Spectronics	Infrared Industries
SPX 2232	7016
SPX XXXX	
	Motorola
RCA	MRD500
C-30807	
C-30831	Quantrad
	003-PIN-T018
Hewlett Packard	
5085-4207	Sharp
	PD-50PI
EGG	
SGD-040-B	
UPT	
PIN-040	
PIN-020	

The nine companies listed essentially have either 1- or 0.5-mm diodes mounted in T018 or T046 hermetic packages. The diode's size is a function of the package and is not optimized for any particular application. If a larger diode is desired, the next size offered is about 2.5 mm in diameter in a T05 package. Another problem is that the mounting plane in either the T046, T018, or T05 package is far from the surface of the package (typically 1.5 to 2.5 mm).

Thus, the devices have not been optimized for proper diameter or mounting plan but only to fit in their packages (T046, T018, or T05). Those packages evolved from the solid-state device market and are not the proper configuration for a PIN photodiode. This results in output coupling losses at least 3 dB higher than necessary.

2.3.4 OPTICAL CONNECTORS

The first concern in the connector area is that the supplier have a military product line. Connector manufacturers that pursue the consumer market do manufacture optical connectors; however, those contacted were unwilling to pursue the military specification market. The connector manufacturers who normally pursue the military market have responded to the fiber-optics market, and the number of manufacturers is increasing but most are currently pursuing single-fiber applications.

Table 2-4 is a representative of materials being offered by optical connector manufacturers.

Table 2-4. Survey of Optical Connectors - January 1978

Company and Model	Type
ITT Cannon	
Unilux	Single
SMA-style	Bundle
Amphenol	
905-119-5006	Bundle (epoxy)
905-119-5014	Bundle (crimp)
906-	Single
Sealectro	
55-907-0149-89	Bundle (SMA)
Prototype	Single in development
Thomas and Betts/Ansley	
998-100	Single
Hughes	
Prototype	Single/bundle
Deutsch	
Prototype	Single/bundle

Most manufacturers address the bundle fiber connector with a connection patterned after an SMA RF coaxial connector. This connector is adequate when using high-loss fiber bundles with many fibers (hundreds).

Connector manufacturers are also addressing the low-loss, single-fiber connector. They have produced various designs for connectors in the 1- to 2-dB loss range when fiber diameter is adequately controlled. However, "adequate control" of fiber diameter can range from no variation in diameter to a few micrometers. Today, most single-fiber connectors cost from 10 to 100 times more than bundle connectors because they are either prototypes or in development.

No manufacturer is addressing a low-count, fiber-bundle connector for medium-loss fibers. Essentially, work has been done at both ends of the spectrum but not in the middle.

2.3.5 *MULTIPORT COUPLERS*

Multiport couplers are truly in the development stage; only four manufacturers would consider supplying such components. Table 2-5 lists these suppliers and describes the coupler type and the number of ports per coupler.

Table 2-5. Survey of Multiport Couplers - January 1978

Company and No. of Ports	Type
Spectronics	
6	Bundle reflective
9	Bundle reflective
16	Bundle reflective
6	Single reflective
9	Single reflective
16	Single reflective
Galileo	
5	Bundle transmissive
9	Bundle reflective
16	Bundle reflective
ITT	
5	Bundle transmissive
9	Bundle transmissive
Hughes	
4	Single transmissive

In all cases, only a few prototype or development couplers have been fabricated. As the table indicates, most work has been done in bundled fibers; activity is just beginning in single-fiber couplers. Both reflective and transmissive couplers have been fabricated.

Apparently, Spectronics has the most experience at coupler fabrication, because they have received various military contract awards in this area. Next in experience is the Galileo Company, with efforts being mainly company funded. Note that fabrication of bundle couplers has been more successful than single-fiber couplers, especially in the area of port-to-port variations.

2.3.6 RECEIVER/TRANSMITTER MODULES

Receiver/transmitter modules have been manufactured by many companies to suit a particular system application being pursued by each company. Table 2-6 lists 15 companies and representative modules they make available.

Table 2-6. Survey of RX/TX Modules - January 1978

Company	Transmitter	Receiver
RCA	C30815	C30847
Texas Instruments	TIXL73	TIXL74
Meret	MLT438	R1101
Devar (Bell & Howell)	529	539
EGG	FOD100	MHZ-018
Radiation Devices	FDT-4-A	FDR-3-A
ITT	2D-TX	2D-RX
Hewlett Packard	Prototype TX	Prototype RX
American Laser Systems	729	728
Sperry Univac	Prototype TX	Prototype RX
Valtec	TTK-D1	TTH-D1
Emtel	6112	6122
Electrophysics	EPH60DA	EPH0DB
General Electric	LE-303	LE-304
Adaptive Systems	161	162

These modules are, without exception, designed for point-to-point links of various lengths, data rates, and complexity and are not applicable to a serial data bus system of the type being sought. No manufacturer of any Rx/Tx module set claims to have a product that will work in a MIL-STD-1553 serial data bus.

Most modules listed are "off-the-shelf," but those listed as prototypes are not. Today, the work done by IBM on the ALOFT program plus further IRAD activities places the internal Rx/Tx capabilities of IBM above those available from external sources.

As of January 1978, it can be said that optical components are still in a state of flux and without standardization. Probably, as system users define their total system goals and determine realistic specifications for components, a more mature line of component products will evolve. Note that the largest projected fiber-optic system user will be in the point-to-point telecommunications market, and this market will greatly affect the components industry.

2.3.7 ANALYSIS OF OPTICAL COMPONENTS

Because the serial data bus application in a military avionic system is a significant departure from a point-to-point telecommunications data link, an analysis was in order. The general parameters of concern were input coupling from source to fibers, connector interface coupling, output coupling from fibers to detector, multiport coupler power division, and total system loss allocation or budgeting.

The analysis was performed by four major programs which extensively investigated all parameters known to affect the optical loss in an optical serial data bus system. The outputs of these programs totally describe the optical performance of the system and hence the requirement of the optical receiver. Sixty-six entries were made in the program, and the performance at each major optical interface (input, connector, and output) is given, plus the overall system results.

2.3.8 THE ANALYSIS PROGRAM - TOLINC

This program fully investigated the coupling of light radiated from a source to a fiber-optic cable. TOLINC's inputs were

- Percent tolerance on diameter
- Core index of refraction
- Fiber diameter (μm)
- Core-to-fiber diameter ratio
- End finish type
- End finish quality
- Source aperture (mm)
- Source cosine pattern exponent
- Number of fibers
- NA
- Off-axis tolerance
- Distance between source and fibers
- Angular displacement of the interface.

TOLINC produced these outputs:

- Average loss (dB)
- Minimum loss (dB)
- Maximum loss (dB).

Table 2-7 lists the results obtained.

Table 2-7. TOLINC Results obtained

Input	Minimum Loss	Typical Loss	Maximum Loss	Limit Loss
Tolerance on fiber diameter (%)	0	2.5	5	10
Core index of refraction	1.452	1.495	1.525	1.615
Fiber diameter (μm)	120	155	215	360
Core-to-fiber diameter ratio	1	0.95	0.9	0.85
End finish type: 1 cleave, 2 polish, 3 coat	1	1	2	3
End finish quality: 1 factory, 2 depot, 3 field	1	2	2	3
Source aperture (mm)	0.8	1	1.2	1.4
Source cosine pattern exponent	40	30	20	10
Number of fibers in bundle	61	37	19	7
NA	0.5	0.3	0.2	0.1
Off-axis tolerance (% of source physical aperture)	1	2.5	5	10
Distance between source & fibers (% of source physical aperture)	0	5	10	20
Angular displacement of interface (^{o})	0	2.5	5	10
Output				
Average loss (dB)	0.700	3.042	9.072	19.703
Minimum loss (dB)	0.700	3.033	9.072	19.703
Maximum loss (dB)	0.700	3.049	9.072	19.703

This data was achieved by fixing all but one of the parameters to the typical loss value and varying the remaining parameter. The resulting data curve would show trends or break points from which the minimum, typical, maximum, and limit values could be determined for that parameter. Once all input parameters were exercised, the cases of minimum, typical, maximum, and limit were run simultaneously to yield loss values at these points.

The results somewhat speak for themselves. Everyone would like to achieve the minimum of only 0.7 dB. More typically, input coupling will be 3 dB. If maximum design limits are imposed, 9 dB would be the worst case, and if tolerances were allowed to extend to limit cases, input coupling could reach values close to 20 dB, showing that this situation is truly out of control.

The analysis also determined the sensitivity of the interface to parameter variation. The program was exercised at typical values, varying each parameter serially. Using minimum values might have caused over emphasis; with maximum values, some variations may not have caused any significant effect.

Following are the results of the variations at the typical point in the order of their sensitivity:

- Source diameter
- Source pattern cosine exponent
- Fiber diameter tolerance
- Distance between source and fibers
- Core diameter-to-fiber diameter ratio
- NA
- Angular displacement
- Off-axis tolerance
- End finish quality
- End finish type
- Index of refraction
- Number of fibers.

The results are not surprising; the size of the source is first, the source cosine exponent is second, and the fiber diameter tolerance is third. This simply shows that the source must be compatible with the fiber-optic cable in size and radiation pattern.

The next group in order of importance includes the distance between source and fibers, core diameter-to-fiber diameter ratio, and NA. This shows that the fiber-cable parameters must be properly related to the source.

The next group in order of importance covers angular displacement, off-axis tolerance, end finish quality, and end finish type. This group shows the importance of control over the interface between the source and bundle, the actual connector parameters, and the termination parameters. Lastly, the index of refraction and number of fibers least affect the input coupling loss.

Simply, the sensitivity of parameters shows that if the source parameters and fiber-cable parameters are compatible, the connector parameters become less important at the input coupling interface.

2.3.9 THE ANALYSIS PROGRAM - TOL4

This program fully investigated the coupling of light from one fiber-optic cable to another at the connector interface.

TOL4's inputs were

- Percent tolerance on diameter
- Core index of refraction
- Core-to-fiber diameter ratio
- End finish type
- End finish quality
- Larger or smaller fiber is the source
- NA
- Off-axis tolerance
- Rotation (degrees)
- Distance between fibers
- Angular displacement of the interface.

TOL4 produced these outputs:

- Average loss (dB)
- Minimum loss (dB)
- Maximum loss (dB).

Table 2-8 lists the results obtained.

This program shows again that when all is minimum, the loss value of 0.6 dB is desirably good. Typically, however, a loss of 1.2 dB can be expected. If maximum design limits are imposed, the worst case would be 2.8 dB maximum. If tolerances were allowed to extend to the limit cases, the connector interface coupling loss could reach a value of 4.8 dB, which cannot be tolerated in a system implementation.

Using typical values as a reference, TOL4 shows the sensitivity of parameters to be in the following order:

- Fiber diameter tolerance
- Core diameter-to-fiber diameter ratio
- NA
- Distance between fiber bundles
- Source larger/smaller
- Rotation
- X/Y shift
- Angular displacement
- End finish quality
- End finish type
- Index of refraction
- Number of fibers.

Table 2-8. TOL4 Results Obtained

Input	Minimum Loss	Typical Loss	Maximum Loss	Limit Loss
Tolerance on fiber diameter (%)	0	2.5	5	10
Core index of refraction	1.452	1.495	1.525	1.615
Core-to-fiber diameter ratio	1	0.95	0.9	.85
End finish type: 1 cleave, 2 polish, 3 coat	1	1	2	3
End finish quality: 1 factory, 2 depot, 3 field	1	2	2	3
Enter 1 if the larger fiber is the source; 0 if the smaller fiber is the source	0	0	1	1
Number of fibers in bundle	61	37	19	7
NA	0.1	0.2	0.3	0.5
Off-axis tolerance (% of source physical aperture)	1	2.5	5	10
Rotation (°)	1	2.5	5	10
Angular displacement of the interface (°)	0	1	3	5
<u>Output</u>				
Average loss (dB)	0.611	1.237	2.762	4.810
Minimum loss (dB)	0.400	1.149	2.661	4.536
Maximum loss (dB)	0.629	1.304	2.959	5.084
<u>Input</u>				
Enter no. of connectors in system	4	4	4	4
<u>Output</u>				
System average	0.608	1.233	2.756	4.812

Note that the fiber diameter tolerance is of first importance; second, core diameter-to-fiber diameter ratio; and third, NA. In other words, the fiber parameters have the most influence on a connector interface.

The next group in order of importance includes distance between fiber bundles, whether the source is the larger or smaller fiber, rotation, X/Y shift, and angular displacement. These parameters are basically controlled by the connector.

End finish type and quality are next in importance and are strictly a function of the terminal terminating processes. Again, the index of refraction and number of fibers least affect connector coupling loss.

In summary, the fiber parameters are most important to control at a connector interface, next are the parameters controlled by the connector, and finally are those parameters concerned with the terminating process.

2.3.10 THE ANALYSIS PROGRAM - TOLOUTC

This program fully investigated the coupling of light from a fiber-optic cable to a photodetector.

TOLOUTC's inputs were

- Percent tolerance on diameter
- Core index of refraction
- Fiber diameter (μm)
- Core-to-fiber diameter ratio
- End finish type
- End finish quality
- Photodiode chip diameter (mm)
- Number of fibers
- NA
- Off-axis tolerance
- Distance between PD chip and fibers
- Angular displacement.

TOLOUTC produced these outputs:

- Average loss (dB)
- Minimum loss (dB)
- Maximum loss (dB)

Table 2-9 lists the results obtained.

Table 2-9. TOLOUTC Results Obtained

Input	Minimum Loss	Typical Loss	Maximum Loss	Limit Loss
Tolerance on fiber diameter (%)	0	2.5	5	10
Core index of refraction	1.452	1.495	1.525	1.615
Fiber diameter (μm)	120	155	215	360
Core-to-fiber diameter ratio	0.85	0.9	0.95	1
End finish type: 1 cleave, 2 polish, 3 coat	1	1	2	3
End finish quality: 1 factory, 2 depot, 3 field	1	2	2	3
Photodiode chip diameter (mm)	2	1.5	1	.5
Number of fibers in bundle	61	37	19	7
NA	0.1	0.2	0.3	0.5
Off-axis tolerance (% of source physical aperture)	1	2.5	5	10
Distance between PD chip & fibers (% of source physical aperture)	50	100	150	250
Angular displacement of interface (^{o})	0	2.5	5	10
Output				
Average loss (dB)	0.250	0.325	3.196	17.049
Minimum loss (dB)	0.250	0.324	3.192	17.049
Maximum loss (dB)	0.250	0.325	3.201	17.049

The program showed that output coupling losses in the minimum case can be extremely low (0.25 dB). Even the typical value can be very low (0.33 dB). If maximum design limits were imposed, about 3.2 dB output coupling loss would result. As a point of interest, this is currently where the very best photodiode would be catagorized.

If tolerances were allowed to extend to the limit case, losses as high as 17 dB could result. Presently, many photodiodes have parameters in this category, mostly because of poor chip packaging (too small a chip, too far from the surface of the package).

Using typical parameters as a reference, TOLOUTC shows the sensitivity of the parameters to be in the following order:

- Photodiode diameter
- NA
- Distance between PD chip and fibers
- Fiber diameter tolerance
- Number of fibers
- End finish quality
- End finish type
- Angular displacement
- Packing fraction
- Off-axis tolerance
- Core index of refraction

Not surprisingly, the size of the photodiode is most important. Next in order of importance are distance between PD chip and fibers, fiber diameter tolerance, fiber diameter, and number of fibers. These parameters are basically controlled by the fiber itself.

End finish type and quality are next, which are basically termination processes. Next is angular displacement, packing fraction, and off-axis tolerance. These are mainly the parameters controlled by the connector, and at the output interface, these parameters are apparently not highly significant. Lastly is core index of refraction, which least affected the output coupling loss.

In review, the photodiode size is most important. Second in importance is the fiber parameters, third is the termination processing, and fourth is the connector parameters.

2.3.11 MULTIPORT COUPLER INFORMATION

To continue with the system analysis, certain information is required on multiport couplers. But, multiport couplers were not analyzed because investigation of this component is still in the research and development stage. Not enough is known to adequately and accurately predict values for excess loss or nonuniformity. Therefore, we used information extracted from the few published reports in this area.

2.3.12 THE ANALYSIS PROGRAM - BUD

This program investigated the loss budgeting of all the individual parameters contributing to the total optical system loss. For each parameter, we assembled either limited analysis data or pertinent information from reports or data sheets.

BUD's inputs were

- LED power - minimum, average, maximum
- Input coupling loss - minimum, average, maximum
- Cable length - minimum, average, maximum
- Loss per unit length - minimum, average, maximum
- Connector loss - minimum, average, maximum
- Number of connectors - minimum, average, maximum
- Number of ports
- Multiport coupler excess loss - minimum, typical, maximum
- Nonuniformity - minimum, typical, maximum
- Output coupling loss - minimum, typical, maximum.

BUD produced these outputs:

- Typical power at detector (dBm)
- Minimum power at detector (dBm)
- Maximum power at detector (dBm)
- Dynamic range (dB).

Essentially, all inputs to the BUD program are known. This program helped us understand the capabilities of high-loss, medium-loss, and low-loss fiber-optic data bus systems. Table 2-10 lists comparative information on those systems utilizing BUD.

Table 2-10. Comparing Systems Utilizing the BUD Program

Input	Low-Loss Fibers			Medium-Loss Fibers			High-Loss Fibers		
	Min.	Avg.	Max.	Min.	Avg.	Max.	Min.	Avg.	Max.
LED output power (dBm)	0	6	10	-5	0	5	-6	0	4
Input coupling loss (dB)	8	9	10	5	6	7	2	3	4
Cable length (km)	0.01	0.1	1	0.001	0.01	.1	0.0001	0.001	0.01
Loss per unit length (dB)	1	2*	4	10	20*	40	100	200*	400
Connector loss (dB)	1	1.4	1.8	1.8	2.2	2.6	2.6	3	3.4
Number of connectors in in system	5	6*	9	2	4*	6	2	3*	5
Number of ports in system	16	16	16	16	16	16	16	16	16
Multiport coupler excess loss (dB)	3	4*	5	2	3*	4	1	2*	3
Nonuniformity (dB)	0	2*	4	0	1*	2	0	.5*	1
Output coupling loss (dB)	1	2*	3	2	3*	4	3	4*	5
Output									
Typical power (dBm)	-31.64			-34.04			-30.74		
Minimum power (dBm)	-54.24			-53.64			-52.04		
Maximum power (dBm)	-19.05			-19.65			-19.25		
Dynamic range (dB)	35.19			33.99			32.79		

* Typical instead of average

In Table 2-10, components for a system in each fiber group have been assembled into a system; "apples and oranges" mixed systems are strictly avoided. Two areas of loss have been kept constant to show the realistic applications areas of the different fiber groups: (1) the loss allocated to the transmission line, and (2) the loss allocated to the connectors in the system.

The first input is the LED power. These numbers represent the range of variation due to device manufacture and ambient temperature. Note that the actual range of the fiber group varies because the device enhancement varies or the actual device type varies.

The input coupling losses are derived by exercising the TOLINC program with sources most compatible with the fiber type. Note the coupling to higher-loss fibers is more effectively accomplished, due mainly to the NA of the fiber group.

The loss for the cable has been kept constant, and therefore the high-loss fiber is utilized in systems having lengths of typically 0.001 km, or 3 ft, with an upper limit of 30 ft and a lower limit of 0.3 ft. The medium-loss fiber is utilized in systems having lengths of typically 0.01 km, or 30 ft, with an upper limit of 300 ft and a lower limit of 3 ft. The low-loss fiber is utilized in systems having lengths of typically 0.1 km, or 300 ft, with an upper limit of 3,000 ft and a lower limit of 30 ft. This is in keeping with the loss parameters shown for high-medium-, and low-loss fibers which show this order of magnitude loss grouping.

The connector losses for the three fiber groups are taken from exercising TOL4. As stated in the review of the TOL4 program, these reflect mainly the parameters of the fiber. Note that the high-loss connectors have proportionally higher losses and the low-loss fibers have lower connector losses. If, as we previously stated, we maintain a fixed allocation for connectors, more connectors can be used in lower-loss systems, as shown.

The number of ports is fixed at 16 in all systems for comparative reasons. The coupler characteristics for excess loss and nonuniformity reflect that high-loss couplers are easier to achieve than low-loss couplers. The output coupling losses are derived from exercising the TOLOUTC program and as previously stated are heavily affected by the fiber characteristics, with the lower-loss fibers having the lower connector loss.

In review, if indeed the fiber loss allocation and connector loss allocation were held constant, a 16-port data bus could be built with any fiber type, provided the number of connectors and length of run limitations shown were adhered to.

It is evident that the system designer can weigh system parameter allocations as he or she wishes. For instance, if fewer ports or connectors were allocated, the increased line lengths could be accommodated.

There are also certain realities which must limit the designer. If we look at the average of the high-, medium-, and low-loss systems, the average allocations are as follows:

Parameter	Loss
LED input power	± 1.5 dBm
Input coupling loss	- 6.0 dB
Cable loss	- 2.3 dB
Connector loss	- 2.2 dB
Number of connectors	4
16-port coupler loss	-12.0 dB
Coupler excess loss	- 3.0 dB
Coupler nonuniformity	± 1.2 dB
Output coupling loss	- 3.0 dB
Typical power to average system	- 33.6 dBm

The same information by component group in order of importance yields

Parameter	Loss
16-port coupler	-15.0 dB
4 connectors	-8.8 dB
Input coupling	-6.0 dB
Output coupling	-3.0 dB
Cable loss	-2.3 dB
LED power	+1.5 dBm

If we look at this information, we realize that the system designer has some freedom, but if he or she is designing a data bus system, there are very definite limitations.

2.3.13 PLASTIC-CLAD, FUSED-SILICA DESIGN CASE

According to our connector (TOL4) analysis, the maximum design limit for core-to-fiber diameter ratio would be 0.9, and the appropriate fiber core diameter for a 19-fiber bundle would be 215 μm. Using these maximum design points, we developed the plastic-clad, fused-silica design case of Table 2-11.

Table 2-11. Design Case for Plastic-Clad Fused Silica

TOLINC	TOLA	TOLOUTC	BUD (Min., Avg./Typ., Max.)
Fiber diameter tolerance (%) 5	Fiber diameter tolerance (%) 5	Fiber diameter tolerance (%) 5	LED output power (dBm) -5 0 5
Core index refraction 1.452	Core index refraction 1.452	Core index refraction 1.452	Input coupling loss (dB) 6.727 6.727 6.727
Fiber diameter (μm) C-to-F ratio 215/0.9	Core-to-fiber diameter ratio 0.9	Fiber diameter (μm)/C-to-F ratio 215/0.9	Cable length (km) 0.001 0.01 0.1
Finish: type/quality 2/2	Finish: type/quality 2/2	Finish: type/quality 2/2	Cable loss (dB) 10 20 40
Source: diameter (mm)/pattern 1.2/20	Source: 1, large; 0, small 1	PD chip diameter (mm) 1	Connector loss (dB) 2.578 2.678 2.875
Number fibers 19	Number fibers 19	Number fibers 19	Number connectors 2 4 6
NA 0.3	NA 0.3	NA 0.3	Number ports 16 16 16
Off-axis tolerance (%) 5	Off-axis tolerance (%)/rotation (°) 5/5	Off-axis tolerance (%) 10	Coupler excess loss (dB) 2 3 4
Distance source-to-fiber (%) 10	Distance fiber-to-fiber (%) 5	Distance fiber-to-PD (%) 150	Nonuniformity (dB) 0 1 2
Angular displacement (°) 5	Angular displacement (°) 3	Angular displacement (°) 5	Output coupling loss (dB) 3.117 3.130 3.152
Average loss (dB) 6.727	Average loss (dB) 2.678	Average loss (dB) 3.130	Typical detector power (dBm) -36.81
Minimum loss (dB) 6.727	Minimum loss (dB) 2.578	Minimum loss (dB) 3.117	Minimum detector power (dBm) -54.17
Maximum loss (dB) 6.727	Maximum loss (dB) 2.875	Maximum loss (dB) 3.152	Maximum detector power (dBm) -24.05
			Dynamic range (dB) 30.12

In the design case, the connector losses are not low enough to ensure that the system will work at the minimum power with a PIN photodectector. But, manipulating the cladding at the connector interface might improve things.

If we removed or stripped the cladding at the interface and close crimped the cores, the core-to-fiber diameter ratio would become 1, the best achievable value. Table 2-12 gives the results of stripping the cladding.

Minimum detector power results are still slightly beyond a silicon PIN photodiode capability. Note that table 2-12 data results from all the parameters in the entire system reaching their design maximum point in the system. This is possible but highly unlikely.

Table 2-12. Design Case for Stripping the Cladding

TOLINC	TOL4	TOLOUTC	BUD (Min., Avg/Type., Max)
Fiber diameter tolerance (%)	Fiber diameter tolerance (%)	Fiber diameter tolerance (%)	LED output power (dBm)
5	5	5	-5 0 5
Core index refraction	Core index refraction	Core index refraction	Input coupling loss (dB)
1.452	1.452	1.452	5.812 5.812 5.812
Fiber diameter (μm)/C-to-F ratio	Core-to-fiber diameter ratio	Fiber diameter (μm)/ C-to-F ratio	Cable length (km)
215/1	1	215/1	0.001 0.01 0.1
Finish: type/quality	Finish: type/quality	*Finish: type/quality*	Cable loss (dB)
2/2	2/2	2/2	10 20 40
Source: diameter/pattern	Source: 1, large; 0, small	PD chip diameter (mm)	Connector loss (dB)
1.2/20	1	1	1. 1.799 1.826 1.926
Number fibers	Number fibers	Number fibers	Number connectors
19	19	19	2 4 6
NA	NA	NA	Number ports
0.3	0.3	0.3	16
Off-axis tolerance (%)	Off-axis tolerance (%)/rotation (°)	Off-axis tolerance (%)	Coupler excess loss (dB)
5	5/5	5	2 3 4
Distance source-to-fiber (%)	Distance fiber-to-fiber (%)	Distance fiber-to-PD (%)	Nonuniformity (dB)
10	5	150	0 1 2
Angular displacement (°)	Angular displacement (°)	Angular displacement (°)	Output coupling loss (dB)
5	3	5	3.185 3.188 3.193
Average loss (dB)	Average loss (dB)	Average loss (dB)	Typical detector power (dBm)
5.812	1.826	3.188	32.55
Minimum loss (dB)	Minimum loss (dB)	Minimum loss (dB)	Minimum detector power (dBm)
5.812	1.779	3.185	-47.60
Maximum loss (dB)	Maximum loss (dB)	Maximum loss (dB)	Maximum detector power (dBm)
5.812	1.926	3.193	-21.61
			Dynamic range (dB)
			26.00

A more realistic representation of the actual worst case would result if we took the system at the typical design values rather than at maximum design values, and let the point at which all these reach their maximum value represent the worst-case system design point.

Table 2-13 gives the results of exercising the analysis program for the typical design case values with stripped cladding. These results shows that a feasible PIN detector system can be built for an aircraft data bus system. The dynamic range, although significantly reduced, is still a rather large value at 23 dB and represents a challenge for the receiver designer.

The preceding design case for plastic-clad, fused-silica fiber represents our recommended approach for the optical hardware in Task Ia of the contract.

Table 2-13. Typical Design Case for Stripped Cladding

TOLINC	TOL4	TOLOUTC	BUD (min., Avg./Typ., Max)
Fiber diameter tolerance (%)	Fiber diameter tolerance (%)	Fiber diameter tolerance (%)	LED output power (dBm)
2.5	2.5	2.5	-5 0 5
Core index refraction	Core index refraction	Core index refraction	Input coupling loss (dB)
1.452	1.452	1.452	5.118 5.118 5.118
Fiber diameter (μm)/C-to-F ratio	Core-to-fiber diameter ratio	Fiber diameter (μm)/C-to-F ratio	Cable length (km)
215/1	1	215/1	0.001 0.01 0.1
Finish: type/quality	Finish: type/quality	Finish: type/quality	Cable loss (dB)
2/1	2/1	2/1	10 20 40
Source: diameter/pattern	Source: 1, large; 0, small	PD chip diameter (mm)	Connector loss (dB)
1.2/20	1	1	1.254 1.256 1.290
Number fibers	Number fibers	Number fibers	Number connectors
19	19	19	2 4 6
NA	NA	NA	NA
0.3	0.3	0.3	16
Off-axis tolerance (%)	Off-axis tolerance (%)/rotation (°)	Off-axis tolerance (%)	Coupler excess loss (dB)
2.5	2.5 2.5	2.5	2 3 4
Distance source-to-fiber (%)	Distance fiber-to-fiber (%)	Distance fiber-to-PD (%)	Nonuniformity (dB)
5	2.5	150	0 1 2
Angular displacement (°)	Angular displacement (°)	Angular displacement (°)	Output coupling loss (dB)
2.5	1	2.5	3.198 3.199 3.200
Average loss (dB)	Average loss (dB)	Average loss (dB)	Typical detector power (dBm)
5.118	1.265	3.199	-29.62
Minimum loss (dB)	Minimum loss (dB)	Minimum loss (dB)	Minimum detector power (dBm)
5.118	1.254	3.198	-43.10
Maximum loss (dB)	Maximum loss (dB)	Maximum loss (dB)	Maximum detector power (dBm)
5.118	1.290	3.200	-19.88
			Dynamic range (dB)
			23.22

2.3.14 OPTICAL SYSTEM SELECTIONS

Using our design case as a reference and the information gathered in the optical components survey, we developed specifications for each system component; the nearest available component of that type is listed in Table 2-14.

Table 2-14. System Component Specifications and Availability

Fiber*
PCS 215 ± 10-μm core, 240 ± 10μm clad, 19-fiber bundle, NA=0.3
Nearest type: Galileo - Gallite "FAT"

Source
Optically enhanced emitter
Nearest type: Spectronics - SPX-XXXX

Connector*
Crimp type for 19-fiber hexagonal PCS
Nearest type: ITT Cannon - SMA TYPE-XXX

Multiport coupler
16-port, 19-fiber bundle, transmissive
Nearest type: Galileo - PTSC - XX-XX

Detector
1-mm PIN, 1-mm max. to surface, isolated, shielded
Nearest type: Spectronics - SPX - XXXX

RX/TX
0 dBm-min. transmitter - 33-dBm typical receiver, 23-dB dynamic range
Nearest type: IBM - modified ALOFT RX/TX

* Substitutes were used in the task Ia hardware.

Many components have parameters close to those needed, but none was "off-the-shelf". For this recommended system, available components would require modifications, but in most cases, these would be minor and not involve large costs or time delays. There was certain exposure in the connector area because no l[illegible]-count, hexagonally packed, and registered connector was available or in design.

In the hardware fabricated for task Ia, problems in terminating PCS fiber made it necessary to use a substitute glass/glass fiber and associated connector.

2.4 OPTICAL MODULATION STUDIES

Various optical waveforms were investigated in an attempt to maximize the signal-to-noise ratio for the system.

Figure 2-2 shows a three-level Manchester waveform as in the 1553A standard, a two-level zero-to-plus Manchester, and a pulse position modulation (PPM), which is easily encoded from, and decoded to a logic level Manchester interface. The PPM is encoded from Manchester by issuing a 50-ns pulse at the beginning of all positive transitions of the Manchester, and again for every 500 ns of "up" level after the first 500. The diagram shows the first few bits of a command word of the 1553 protocol, encoded in the three forms. Both the three- and two-level Manchester were rejected in favor of the pulse modulation because of the gains achievable.

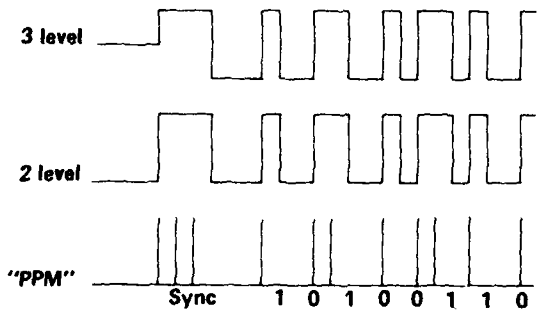


Figure 2-2. Manchester Waveforms

The third ("off") state of the three level waveform above was handled by a 2-μs timeout in the PPM receiver decode logic. Because no pulses can be farther apart than 1.5 μs, a detection of the timeout after any pulse signalled the end of a message.

PPM has three potential advantages: (1) a gain in S/N ratio proportional to the square root of the ratio of the relative amplitude of the short pulse to the amplitude of a conventional average 50% duty cycle Manchester, (2) a constant, nonvariable threshold level at the decision point for "1s" and "0s" in the receiver, and (3) a simple way to handle the dynamic range of detector input power caused by differences in highest- and lowest-loss paths between terminals on the bus.

For the following derivations and for a first-order comparison with other systems, two assumptions will be made: (1) the principal source of noise is in the receiver preamplifier, and (2) the noise is white Gaussian in nature. These assumptions are compatible with previous literature on the subject and are also real in terms of the systems and bandwidths involved.

Figure 2-3 shows the type of pulse examined, along with definitions of amplitude and normalized time. Figure 2-4 shows the receiver bandpass characteristics, along with a low-frequency, f_L, and high-frequency, f_H, cutoff at the 3-dB and 3- or 6-dB points, respectively.

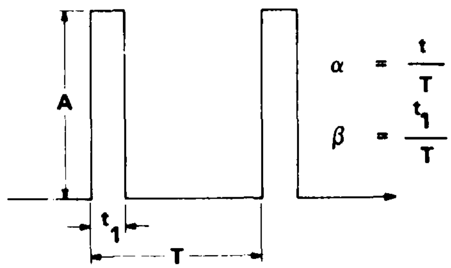
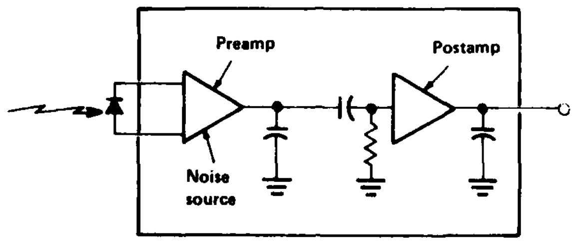


Figure 2-3. Type of Pulse Examined



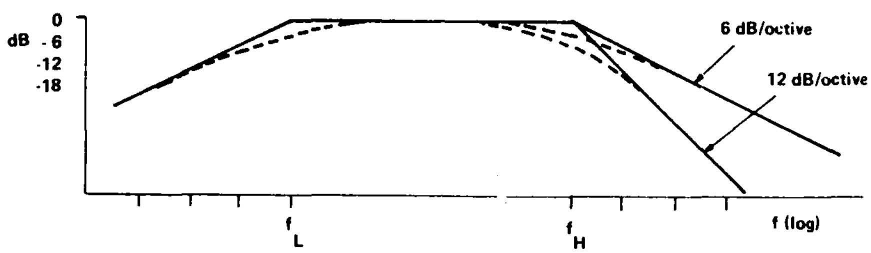


Figure 2-4. Receive Bandpass Characteristics

The autocorrelation function for Gaussian white noise after the receiver "filter" is

$$R_O (t) = \emptyset \int_0^{\infty} h(u) \cdot h\left(u + |t|\right) du$$

where h is the receiver impulse response, and $\emptyset$ is the noise power per unit frequency.

From this it follows that the standard deviation in rms noise is

$$\sigma = \sqrt{\emptyset \int_0^{\infty} [h(u)]^2 du}$$

= rms noise at receiver output.

A figure-of-merit (FOM) for this pulse system can be chosen as

$$\text{FOM} = \frac{A'}{\sigma} \quad \text{(Signal-to-rms-noise ratio)}$$

where A' is the pulse amplitude at the receiver's output.

By substituting for σ, multiplying numerator and denominator by A $\sqrt{t_1}$, and substituting t for u,

$$\text{FOM} = \frac{\sqrt{A} \quad \sqrt{At_1}}{\sqrt{\phi}} \cdot \frac{A'}{A\sqrt{t_1 \int_0^\infty [h(t)]^2 dt}}$$

where A is the input pulse amptitude and t_1 is the pulse width.

To maintain a relatively constant temperature rise in the LED junction, the product At_1 must be kept constant. Because for a given receiver, ϕ is a constant, we can write the last equation as

$$\text{FOM} = \sqrt{A} \cdot C \cdot A_n$$

where $$A_n = \frac{A'}{A\sqrt{t_1 \int_0^\infty [h(t)]^2 dt}} \quad \text{(normalized output} \qquad 1)$$

and $$C = \sqrt{\frac{At_1}{\phi}}$$

It follows, then, that the FOM is a function of the receiver bandpass characteristics, the pulse width, and the input amplitude. The significant factor here is the gain in FOM by the square root of the input pulse amplitude. Also, for a given pulse width and duty cycle, there is an optimum bandpass with a precisely located upper and lower cutoff frequency to give the maximum FOM. This is basically the so-called matched filter.

A computer program was written whereby if a filter bandpass was given (f_H-f_L), the f_H was selected for giving the greatest peak value of A_n for a given input pulse. This bandpass was then varied to find the bandpass and upper frequency location which would give the absolute maximum peak for A_n. All of the output pulses are for a steady-state condition after a long train of pulses 500 ns apart. This was selected as the worst case for the peak signal amplitude above some previously defined threshold.

Figure 2-5 shows the resultant shape of a pulse like Figure 2-3 for $\beta = 0.1$, after passing through a filter with characteristic as shown in Figure 2-4 (12 dB/octave upper cutoff). The solid curve in Figure 2-5 is the value of A_n over the pulse period, and the dotted curve is the value A'/A over the period. This particular curve is for the optimum FOM and occurs for values of f_L and f_H of 1 MHz and 8 MHz, respectively. The peak valve of the solid curve would be the optimum FOM, in this case, 0.85.

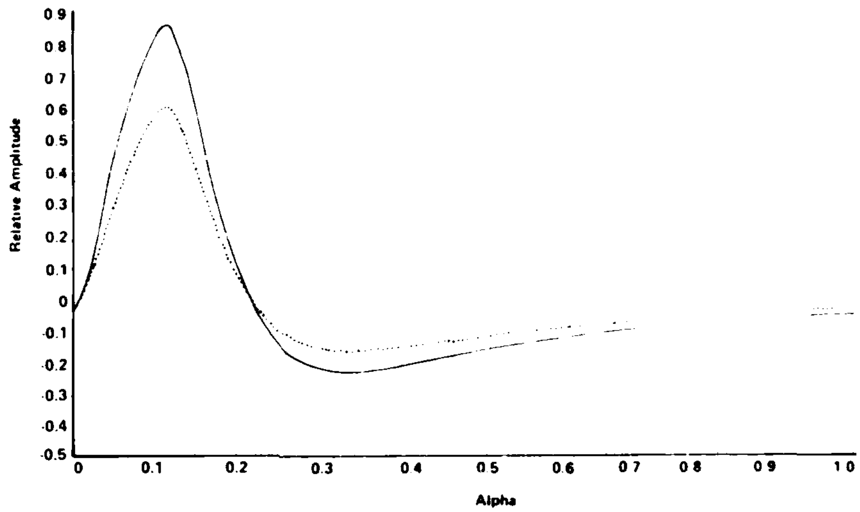


Figure 2-5. Normalized Pulse Response

To reconstruct the logic-level Manchester-encoded signal, the pulse must activate a logic gate when it exceeds some threshold, and deactivate it when it falls below this threshold. This gate must not be activated by noise between pulses. Figure 2-6 shows this concept (its error probability is analyzed later). In the figure, dimension A is the signal

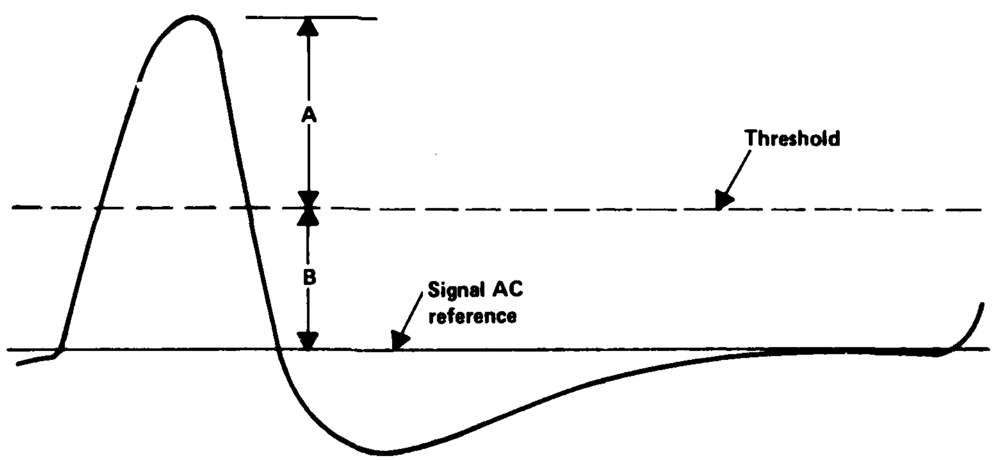


Figure 2-6. Manchester-Encoded Signal

amplitude above noise required to activate the circuit with a given error probability, and dimension B is the threshold amplitude above the AC reference to prevent a false alarm with a given probability.

When comparing a normal 50% average duty cycle Manchester with the PPM system, using a 50-ns pulsewidth for a 1-MHz bit rate, we find the term A_n for both systems, when optimized for bandwidth, is about the same ($\simeq$ 0.85). However, with a 50-ns pulse, one can allow an increase of 10 times the amplitude for a factor of $\sqrt{10}$, or +5 dB of peak power or peak S/N ratio. For LEDs, in their normal operating range, this requires a 1-A pulse 50 ns wide.

Pulse widths much less than 50 ns require very fast LEDs and correspondingly more difficult peak currents to generate. The pulse was therefore limited in this study to 50 ns and 1 A. Dr. J.R. Biard of Spectronics has performed an analysis in a paper ("Thermal Transients in Light-Emitting Diodes") which shows that 1-A, 50-ns, 5% duty cycle pulses maintain junction temperature rises which will preserve conditions that will not degrade the life of the device analyzed.

An advantage of PPM not mentioned before is the compatibility with a natural operating mode of lasers. When semiconductor laser technology has matured, the emitter can be easily replaced by a laser with minimum modifications to the system and gain additional dB through the $\sqrt{A}$ factor.

A further advantage, as previously stated, is that with the bandwidths involved, the receiver has essentially recovered between pulses such that a changing threshold for I/O decisions is not required because of the dynamic range, as is required in the 50% duty cycle case. One sets the threshold for the minimum signal and leaves it there.

A related advantage is the method for handling the large dynamic signal range. To adjust the threshold levels in the 50% duty cycle case, a fast AGC system is required because the threshold must be set in the middle of the signal amplitude. In the case of the short pulse, the threshold is constant, and one needs only to compress the high signal levels with techniques such as logarithmic amplifiers or networks.

In the PPM system, we are concerned with two types of errors: the probability of a missed detection and the probability of a false alarm. The differential false alarm probability, dp, that the noise exceeds some threshold, $k\sigma$ (σ = rms noise), between a time t and (t + dt) is

$$dp = \varepsilon^{-\frac{1}{2}k^2} \bar{f}\, dt$$

where $\bar{f}$ is defined as the average frequency of receiver bandpass noise spectrum and is given by

$$\bar{f} = \sqrt{\frac{\int_{-\infty}^{\infty} f^2 \phi_O \;(f)\, df}{\int_{-\infty}^{\infty} \phi_O \;(f)\, df}} = \frac{1}{2\pi T} \sqrt{\frac{\int_0^{\infty} [h'\,(t)\,]^2\, dt}{\int_0^{\infty} [h\,(t)\,]^2\, dt}}$$

where ϕ_O (f) is the Fourier transform of the noise autocorrelation function Ro (t) given before and is therefore the spectrum of noise in the filter output, and where h (t) is the impulse response of the receiver, and h' (t) is its first derivative.

The average number of threshold crossings, P, in time t is

$$P = \int_0^t \bar{f}\, \varepsilon^{-\frac{1}{2}k^2}\, dt$$

If we normalize time as in Figure 2-5. and define a term $\bar{x} = 2\pi \bar{f}T$, where T is the reciprocal of the average pulse repition frequency, then

$$P = \int_0^{\alpha} \frac{\bar{x}}{2\pi} \varepsilon^{-\frac{1}{2}k^2}\, d\alpha$$

$$\text{and } P = \frac{\bar{x}}{2\pi}\, \varepsilon^{-\frac{1}{2}k^2} \qquad \text{for } \alpha = 1$$

The equation for P describing the average number of threshold crossings in time t did not limit the duration of the threshold crossing. Because the 1/0 decision in reality is a function of the electronic response of the decision-making circuit, the noise must remain some minimum time, t_o, above the threshold for a false alarm to occur. The true false alarm probability, P_{fa}, is therefore

$$P_{fa} = P' \cdot P = P' \cdot \frac{\bar{x}}{2\pi} \varepsilon^{-\frac{1}{2}k^2}$$

where P' is the conditional probability that if the noise exceeds the threshold $k\sigma$, it will stay above it for at least time t_o. P' is given as

$$P' \cong \varepsilon^{-\frac{1}{2}(\pi \bar{f} t_o k)^2} = \varepsilon^{-\frac{1}{2}(\bar{x} t_o k/2T)^2} \qquad [(1)]$$

[(1)] Stationary and Stochastic Processes, Cramer & Leadbetter

The false alarm probability is therefore

$$P_{fa} \cong \frac{\bar{x}}{2\pi} \varepsilon^{-\frac{1}{2} k^2 [1 + (\bar{x} t_o/2T)^2]}$$

Or, for a given P_{fa}, the threshold, $k\sigma$, must be set at

$$k\sigma = \sigma \sqrt{\frac{2(\ln \frac{\bar{x}}{2\pi} - \ln P_{fa})}{1 + \left(\frac{\bar{x}\ to}{2T}\right)^2}}$$

The probability of a missed detection, P_{md}', is the Probability that $[A' + n(t)] < k\sigma$ for some time t in the interval t_o. Because n (t) is the Gaussian variable with standard deviation σ, then

$$P_{md} \leq \int_{\Delta}^{\infty} \varepsilon^{-\frac{1}{2} u^2} du$$

where $\Delta = \frac{1}{\sigma}(A' - k\sigma)$

This is the well-known error function, erfc.

In a 1553 bus system using PPM, the average pulse spacing during a transmission is 1 μs. Under these conditions with a pulse width of 50 ns, the bandwidth and upper and lower cutoff frequencies are 8.4, 8.8, and 0.4 MHz, respectively. From the previous equations for P_{fa} and P_{md}, if it is desired that the BER should be 10^{-12}, then the threshold, $k\sigma$, must be 7.15σ for false alarms, and the pulse amplitude, A', must be $7.03\sigma + 7.15\sigma = 14.18\sigma$. This is a limiting value because false alarms do not occur during pulses, and pulses have finite widths greater than the response times of the decision circuitry. It is additionally pessimistic because it does not take into account detection of errors by parity.

The threshold and relative pulse amplitudes for a 50% duty cycle straight Manchester are about the same. However, note that the PPM pulse using LEDs can achieve 10 times the amplitude and therefore have an advantage of $\sqrt{10}$, or +5 dB in S/N ratio and +10 dB in terms of amplitude.

More details on the various parameter tradeoffs can be found in the Task I Analysis Review Report — Contract Data Item A005.

The foregoing analysis assumed the noise was Gaussian (noise power/Hz independent of frequency). For a first-order waveform comparison, this is an adequate assumption. It also assumes that the sampling point for determing "1s" and "0s" on the reconstructed waveform is at the peak of the amplitude.

Because the Manchester-encoded waveforms (full width or PPM) contain their own clock information, this sampling time is determined by synchronizing with a waveform transition during the invalid manchester-encoded sync time of the Manchester-encoded word. In the PPM modulation, this time can not be ambiguous by more than $\pm$25 ns; nor can the transitions of the reconstructed logic level signal be off by more than ± 25 ns

It is evident, then, that the sampling time for each bit in this system will always be within the limits of the reconstructed waveform. This is not nececessarily true for full-width Manchester and is a function of the receiver bandwidth.

An analysis compared PPM to full-width Manchester when taking into account non-Gaussion noise and also noninfinite receiver bandwidths. Another factor, waveform transition jitter, was also accounted for. The results of this analysis are shown in Table 2-15 in terms of additional attenuation allowable in the bus terminal-to-terminal transmission path for a PPM system to achieve a 1×10^{-9} bit error rate. This analysis is detailed in Appendix A.

Table 2-15. Additional Attenuation Allowable for PPM (dB)

Jitter (ns)	Full-Width Receiver Cutoff Frequency (Hz)		
	1.5×10^6	2×10^6	3×10^6
± 25	4.3	4.7	6.0
± 50	4.7	5.1	6.2

2.5 RECOMMENDED SYSTEM

As a result of the studies described in 2.3 and 2.4, IBM recommended the system shown in Figure 2-7. The details or design features recommended are in Table 2-16. In Figure 2-7, only two of a possible 16 terminals are shown.

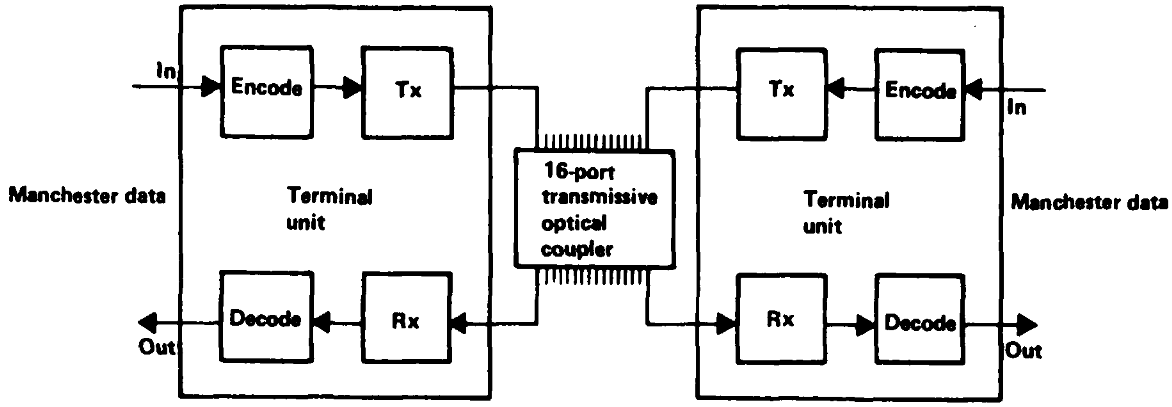


Figure 2-7. Recommended System

Table 2-16. Recommended Design Features

• Source:	LED (SE2231)
• Fiber:	215-μm core PCS
• Cable:	19 fibers in ESM heavy-duty cable
• Connectors:	Hexagonally packed, stripped, and thinly reclad fibers with keyed connection
• Coupler:	16-port transmissive multiport coupler
• Detector:	PIN photodiode (SD3478)
• Optical waveform:	PPM Manchester
• Transmitter:	TTL level input, switch driver, 1-A switch, drive supply, drive store
• Receiver	Transimpedance preamp, postamp, and two stages of amplification and dynamic range compression into a high-speed comparator with TTL levels out

2.6 TEST PLAN

At the Task I Analysis Review, a Task Ia development and test plan was presented. Before demonstration in a system, three levels of test were proposed: component, subsystem, and link. Detailed descriptions of the proposed tests may be found in Contract Data Item A004, Task Ia Development and Test Plan.

All components purchased would be tested to their vendor's data sheets or IBM's procurement specification. The next series

of tests would be at the subsystem level; that is, transmitters, receivers, encode/decode logic, etc. Operation as a terminal-to-terminal link would be verified before installation in the demonstration system.

The demonstration system proposal was essentially a mockup of the A-7 navigation, bombing, and weapon delivery electronics. IBM had already installed two redundant 1553A-compatible twisted, shielded-pair wire buses, one of which would be replaced by the proposed fiber-optic bus. The remaining wire bus and the fiber-optic bus would be compared in terms of system operation, and word and bit error rates. Objectives of the test would be proving equal or better system performance with the fiber-optic bus.

Section 3
TASK Ia DETAILS

This section describes the hardware design, final hardware definition, test configurations, and test results of Task Ia.

3.1 HARDWARE DESIGNS

3.1.1 TRANSMITTER

A Spectronics 2231 LED emitting in the infrared range at 907 nm was recommended as the optical source for the fiber-optic bus. Tables 3-1 and 3-2 list the physical and optical/electrical characteristics of the device as published by the vendor.

Table 3-1. SPX 2231 LED Physical Characteristics

1. Designed for coupling to 0.045- to 0.050-inch diameter fiber-optic bundles.
2. TO-46 package configuration.
3. Thermal resistance 125°C/W (typical).
4. LED isolated from case. Case may be grounded for EMI shielding.
5. Planar GaAs edge-emitting wafer with optimized reflector for high efficiency.
6. Operating temperature -65°C to +125°C.
7. Case soldering temperature 240°C maximum for 3 minutes.

Table 3-2. SPX 2231 LED Optical/Electrical Characteristics (25°C)

1.	BV_R	=	4.0 V (max) @ I_R = 10 μ A
2.	V_F	=	1.5 V (max) @ I_F = 100 mA
3.	r_s	=	1.0 Ω (typ)
4.	C_T	=	60 pF (max) @ V_R = 1 V, f = 1 MHz
5.	λ p	=	907 nm @ I_F = 100 mA
6.	$\Delta\lambda$	=	24 nm @ I_F = 100 mA
7.	P_O	=	2.0 mW (min) @ I_F = 100 mA
8.	θ	=	15°, cone half angle for 1 mW
9.	t_r	=	20 ns (typ) for current pulse.

Section 2.4 previously described the optical modulation scheme for which the transmitter was designed. A current pulse of 1A with a duration of 50 ns and an average duty cycle of 5% is required to drive the LED in the Pulse Position Modulation mode.

An ideal device for such an application is an enhancement-mode VMOS power FET. It has no storage delay time and allows current rise and fall times typically in the order of 4 ns. The device chosen was a Siliconix VN33AK. Its electrical characteristics are listed in Table 3-3.

Table 3-3. Siliconix VN33AK Electrical Characteristics

					Min.	Typ.	Max.	Min.	Typ.	Max.	Min.	Typ.	Max.			
1	STATIC	BV_{DSS}	Drain-Source Breakdown		35			60			90			V	V_{GS} 0, I_D 10 A	
2		$V_{GS(th)}$	Gate-Threshold Voltage		0.8		2.0	0.8		2.0	0.8		2.0		V_{DS} V_{GS}, I_D 1 mA	
3		I_{GSS}	Gate-Body Leakage			0.5	100		0.5	100		0.5	100	nA	V_{GS} 15 V, V_{DS} 0,	
4							500			500			500		V_{GS} 15 V, V_{DS} 0, T_A 125°C (Note 2)	
5		I_{DSS}	Zero Gate Voltage Drain Current				10			10			10	μA	V_{DS} max. rating V_{GS} 0	
6							500			500			500		V_{DS} 0.8 max. rating, $V_{GS}=0$, T_A 125°C (Note 2)	
7						100			100			100		nA	V_{DS} 25 V, V_{GS} 0	
8		$I_{D(on)}$	ON State Drain Current		1.0	2.0		1.0	2.0		1.0	2.0		A	V_{DS} 25V, V_{GS} 10 V	(Note 1)
9		$V_{DS(on)}$	Drain-Source Saturation Voltage	VN33AK VN66AK VN98AK		1.0			1.0					V	V_{GS} 5 V, I_D 0.3 A	
10							1.8			3.0			4.0		V_{GS} 10 V, I_D 10 A	
11				VN35AK VN67AK VN99AK		1.0			1.1			1.2			V_{GS} 5 V, I_D 0.3 A	
12							2.5			3.5			4.5		V_{GS} 10 V, I_D 10 A	
13	DYNAMIC	g_{fs}	Forward Transconductance		170	250		170	250		170	250		mʊ	V_{DS} 24 V, I_D 0.5 A	
14		C_{iss}	Input Capacitance			[illegible]	40		[illegible]	40		[illegible]	40	pF	V_{GS} 0, V_{DS} 24 V, f 1.0 MHz (Note 2)	
15		C_{oss}	Common Source Output Capacitance			[illegible]	45		[illegible]	40		[illegible]	40			
16		C_{rss}	Reverse Transfer Capacitance			7	10		6	10		5	10			
17		t_{on}	Turn On Time			[illegible]	8		[illegible]	8		3	8	ns	See switching time test circuit VNAR, Section 3	(Note 2)
18		t_{off}	Turn Off Time			[illegible]	8		[illegible]	8		[illegible]	8			

VNAR

Notes:

1. Pulse test - 80 - μs pulse, 1% duty cycle.
2. Sample test.

Absolute Maximum Ratings:

Maximum drain-source voltage
- VN33AK, VN35AK . . . . 35 V
- VN66AK, VN67AK . . . . 60 V
- VN98K, VN99AK . . . . 90 V

Maximum drain-gate voltage
- VN33AK, VN35AK . . . . 35 V
- VN66AK, VN67AK . . . . 60 V
- VN98AK, VN99AK . . . . 90 V

Maximum continuous drain current . . . . 2.0 A
Maximum pulsed drain current . . . . 3.0 A
Maximum forward gate-source voltage . . . . 30 V
Maximum reverse gate-source voltage . . . . 30 V
Maximum dissipation at 25°C case temperature . . . . . . 6.25 W
Linear derating factor . . . . 50 mW/°C
Temperature (operating and storage) . . . . -55 to +150°C
Lead temperature
(1/16 inch from case for 10 seconds) . . . . 300°C

Figure 3-1 is a schematic of the transmitter circuit. Transistor groups U1 and U2 are chips each containing five transistors (3045s). Transistor group U1, comprising a "totem pole" amplifier driven by a negative-going, TTL-level, 50-ns pulse, supplies a voltage to the gate of the VMOS power FET, turning it on, thereby allowing the charge on capacitors C7 and C8 to supply a current pulse through the LED. The time constant of the current path from the capacitor through the LED and FET is small enough to allow a current rise time of 5 ns to an amplitude of 1 A.

Transistors in the U2 group, pins 1 through supply current to charge capacitors C7 and C8 between pulses. This current is equal to the average LED current, which is 1 A times the average duty cycle of 5%, or 50 mA. Resistor R7 allows a small quiescent current to flow through the LED and enhances the light pulse rise time. The remaining transistors in group U2 form a temperature-compensating network with RT1 (thermistor) to control the charge on capacitors C7 and C8, which in turn controls the current through the LED and maintains a peak light output that varies minimally with temperature. Except for the LED, the entire circuit is a hybrid contained in a module as described in Section 3.2.

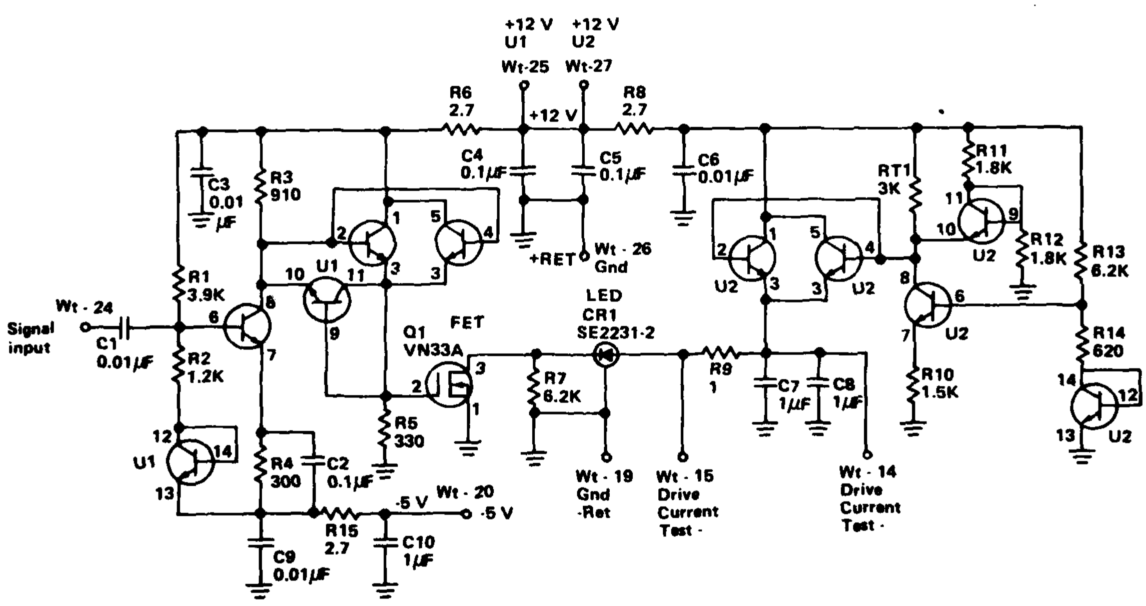


Figure 3-1. Transmitter Schematic Design

3.1.2 RECEIVER CIRCUIT

The recommended optical detector for the bus is a Spectronics 3478 PIN photodiode. Tables 3-4 and 3-5 list the physical and optical/electrical characteristics of the device as published by the vendor.

The function of the receiver is to amplify the output current of the photodiode and reproduce, at TTL levels, the input pulse to the transmitter over a wide range within a given bit error rate (BER). Section 2.4 discussed the bandwidth and signal-to-noise ratios necessary for 1×10^{-12} BER when working with minimum signal. Figure 3-2 is a schematic diagram of the receiver circuit. As in the transmitter circuit, the transistors used are type 3045 (five on a chip). The receiver, except for the photodiode and comparator, is also a hybrid and is described in Section 3.2.

Transistors U1, pins 6 through 10, are configured as a transimpedance amplifier with a transfer characteristic of 4,300 V/A.

Table 3-4. Spectronics 3478 Physical Characteristics

1. Designed for coupling to fiber-optic bundles, active area diameter 0.050 inch.
2. TO-46 package configuration.
3. Detector element isolated from case. Case may be grounded for EMI shielding.
4. Operating temperature -65°C to +125°C.
5. Case soldering temperature 240°C maximum for 3 minutes.

Table 3-5. Spectronics 3478 Optical/Electrical Characteristics (T case = 25°C)

Parameter	Test Condition	Symbol	Parameter Value Min.	Typ.	Max.	Units
Peak response wavelength		λ_p		907		nm
Flux responsivity	Note 1	R		0.5		A/W
Dark current	V_n = 100 V	I_o		1		nA
Response time	V_n = 15 V	t_r		5.0		ns
	V_n = 90 V			1.1		
Series resistance		r_s		15		Ω
Total capacitance	V_n = 20 V					pF
	Anode grounded	C_A		4.4		
	Cathode grounded	C_C		3.0		
Field of view	Note 2			70		degrees
Leakage temperature coefficient		$\frac{\Delta I_o / I_o}{\Delta T}$		10		%/°C

Notes:

1. Measured with a 1.14-mm diameter, 0.66-nA fiber-optic bundle centered on optical axis
2. Angle between 50% response points measured with constant irradiance over an area greater than the area of the optical aperture.

It converts the photodiode current to a voltage that drives an inverter (U1, pins 1 through 5). This voltage, which increases positively with increases in light detected at the photodiode, is then amplified nonlinearly by three successive differential amplifier stages and an emitter follower (U4 pins 1 through 5).

As shown in 2.4, the receiver bandpass characteristics must be 3 dB down at 1 MHz and 6 dB down at 8 MHz, with a 6-dB/oct. and 12-dB/oct. rolloff, respectively. The low-frequency cutoff is accomplished by the two RC network C10, R24. The high-frequency cutoff is accomplished by the two RC networks C8, R19 and the collector capacitance of U4 pin 11, plus C6, R33, and the collector capacitance of U3 pin 11.

Up to the input side of C10, the amplifier is DC-coupled with a very low pass feedback path from U4 pin 3 to U2 pin 9. This stabilizes the operative points of all the differential amplifiers.

The dynamic range of input signal is controlled by the differential amplifiers in that they are powered by current sources at their emitter point, and the current through either side (and therefore the change in collector voltage) cannot exceed this value, regardless of the voltage at the input transistor base.

The input signal to the comparator (U5, pin 1) is as shown in Figure 2-6, and the output (U5, pin 5) is a TTL-level pulse whose width is determined by the positive- and negative-going crossing times of the threshold (U5, pin 2) by the input.

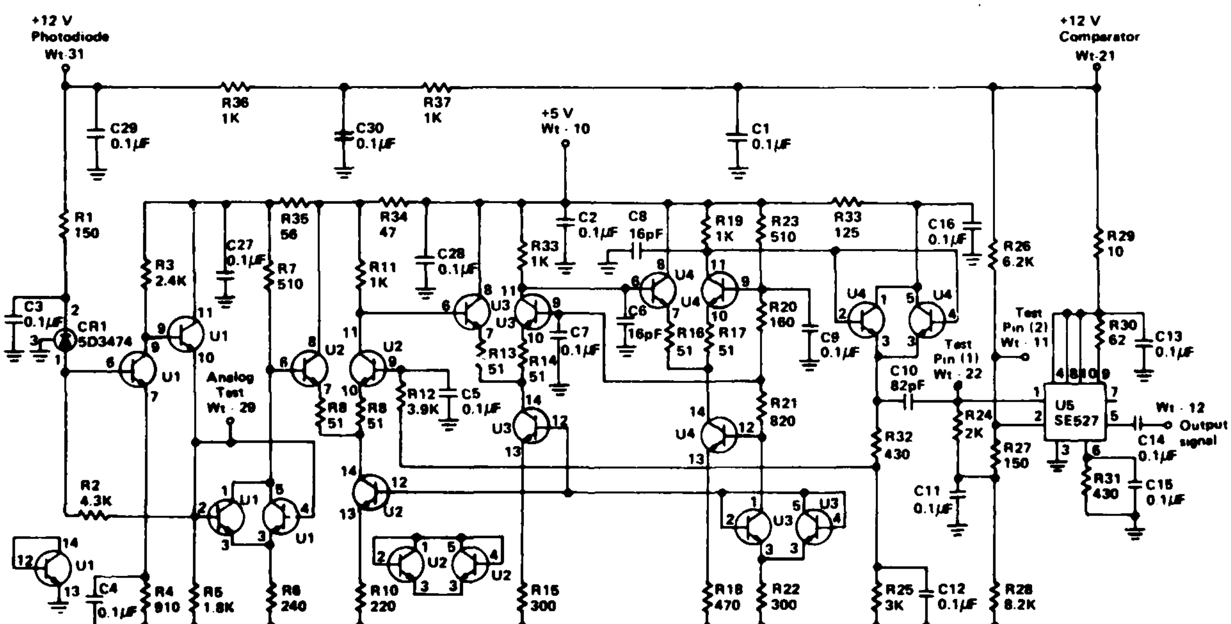


Figure 3-2. Receiver Schematic Diagram

3.1.3 ENCODER CIRCUIT

Figure 2-7 showed a terminal unit consisting of an encoder, a transmitter, a receiver, and a decoder. The encode/decode logic interface consists of a TTL level standard Manchester encoded input and a similar output. The encoder, Figure 3-3, samples the Manchester signal every 500 ns. If the sampled waveform is a logic "1", a 50-ns pulse is issued to the LED transmitter circuit. No pulse is issued for a logic "0". The Manchester data are thus converted to a TTL level pulse position modulated (PPM) waveform. These 50-ns pulses are translated into 50-ns optical pulses by the transmitter.

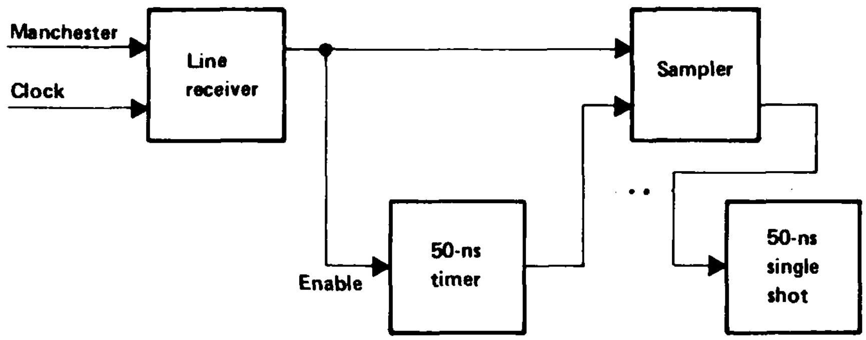


Figure 3-3. Encoder Logic Diagram

3.1.4 DECODER CIRCUIT

The decoder logic converts a PPM-encoded Manchester signal as described in Section 3.1.3 to a standard Manchester-encoded TTL level signal. The decoding logic, Figure 3-4, gives the following three 1553 signals as outputs:

- Zero Crossing Data - Determine polarity of data
- Minus Threshold - For Manchester 1553 systems that combine positive and negative threshold detection
- Plus Threshold - Active whenever the Zero Crossing Data output is active until 2 ns after the last negative transition.

Figure 3-5 shows a sample Manchester waveform input to an encoder and its subsequent reconstruction at the decoder outputs.

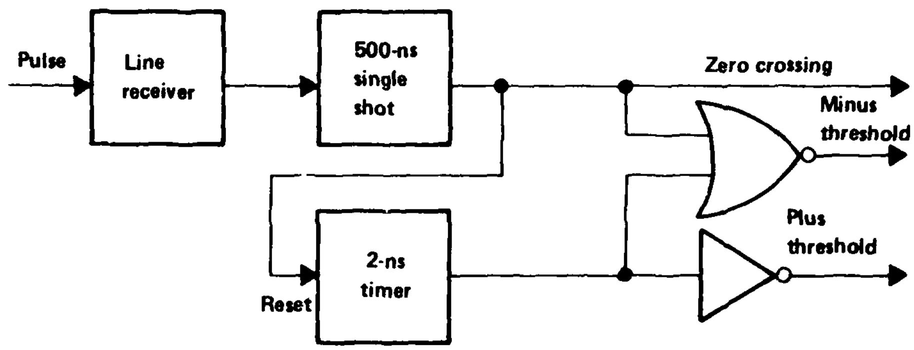


Figure 3-4. Decoder Logic Diagram

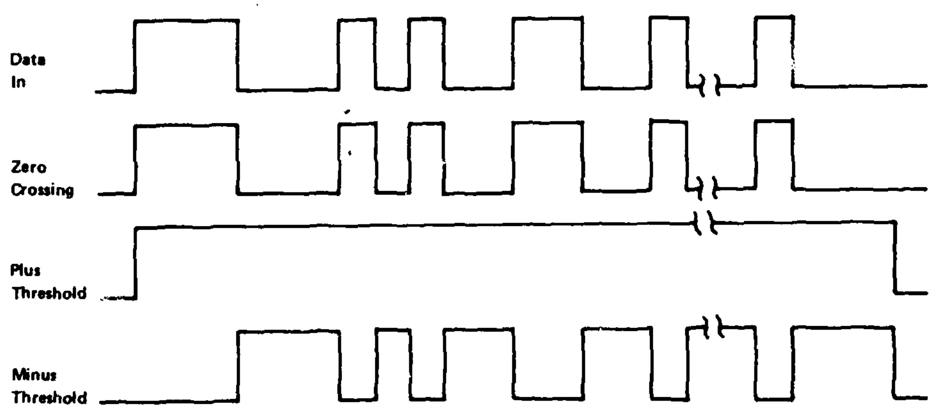


Figure 3-5. Decode Inputs and Outputs

3.2 TASK IA IMPLEMENTED HARDWARE DESCRIPTION

3.2.1 CABLE DESCRIPTION

- A 19-fiber stripped and reclad closepack termination could not be performed with the new connectors. All mechanical tolerances were met, but a material suitable for the strip and reclad operation could not be found. The terminal alignment was correct, but the

fiber core/clad interface was destroyed by recladding, resulting in extremely high losses (9 dB or more).

As an alternative, IBM purchased a substitute high-loss fiber cable and a connector suitable for use with this type of fiber. The fiber used was a Gallite 2000, in a 0.045-inch diameter bundle. The connector was an Amphenol 905 series.

If the same cable loss was allocated to the fiber cable, the run length must be limited to 0.01 km (30 ft) and the number of connectors limited to four. These restrictions did not impact the implementation of the demonstration system in the A-7 Avionic Integration Laboratory because equipment were not separated by more than 30 ft. None of the other system components or their specifications was affected by this cable/connector substitution.

3.2.2 RECEIVER/TRANSMITTER DESCRIPTION

The system specifications for the optical receiver and transmitter led to the construction of two hybrid modules, shown in Figure 3-6 in various stages of assembly. Both units used thin-film hybrid technology and integrated transistor arrays. The electro-optic devices (LED, PD) were included in the package and interfaced by an optical connector.

The receiver, Figure 3-7, was the basis for choosing thin-film hybrid technology because of its low-noise characteristics and good resistor-tracking properties. The package could have been smaller, but it was cost-effective to use a standard package that was immediately available. The receiver met all performance criteria required by the system and became an excellent building block in the overall terminal unit design.

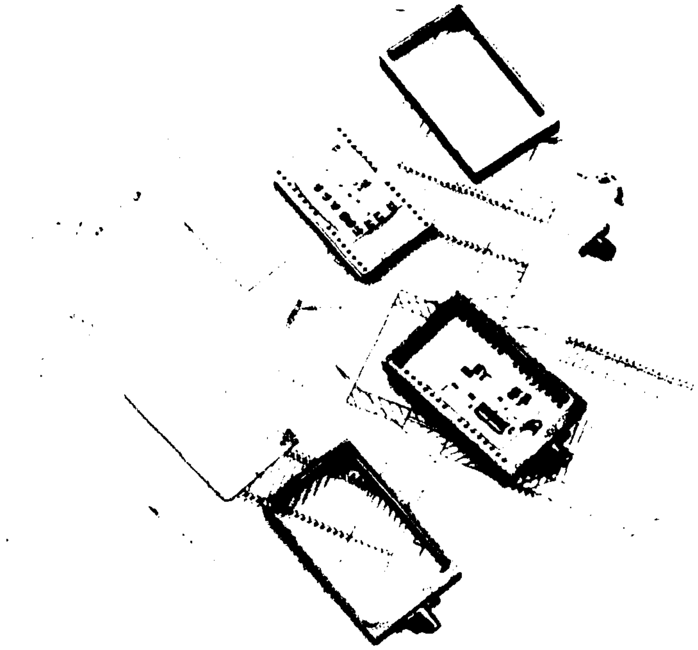


Figure 3-6. Hybrid Modules

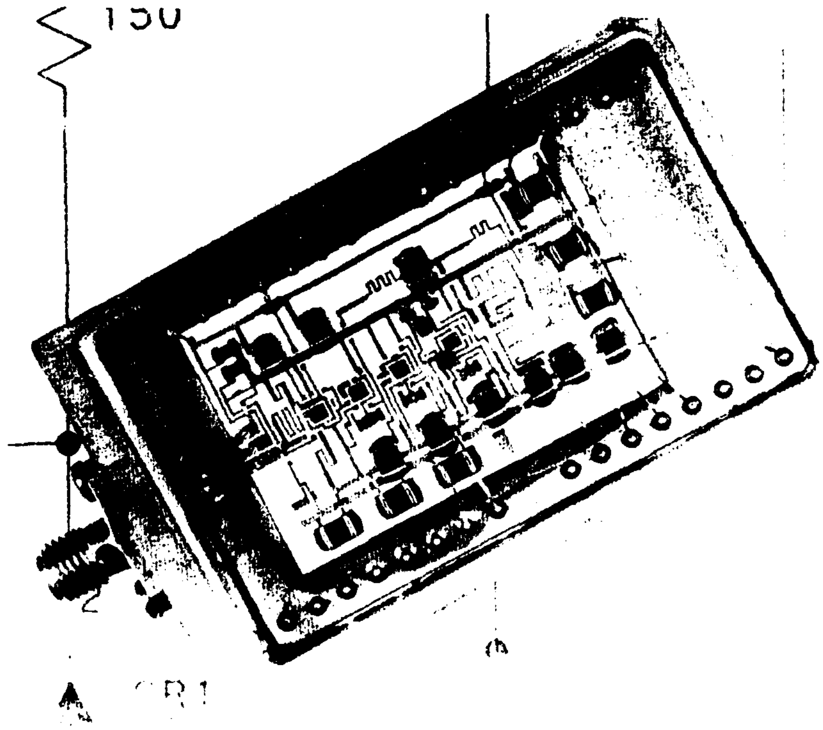


Figure 3-7. Receiver Module

The transmitter, Figure 3-8, was also built in thin-film technology. The package was the same as the receiver's for the sake of commonality and cost effectiveness. This module integrated thermistors, resistors, transistor arrays, large-value capacitors, and VMOS FET transistors in a single package along with the LED. This hybrid produced the temperature-compensated, high-current drive required by the LED, which generated the required optical signal.

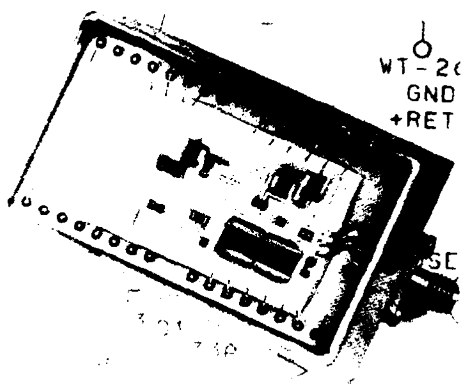


Figure 3-8. Transmitter Module

3.2.3 LOGIC DESCRIPTION

The encode/decode logic function and the clock oscillator were assembled using dual-in-line (DIP) packages mounted on an EECO half-frame wirewrap board. The two hybrids (receiver/transmitter) were also mounted on a printed circuit card the size of an EECO half frame. Both boards were then mounted in a standard single-frame EECO package, as shown in Figure 3-9.

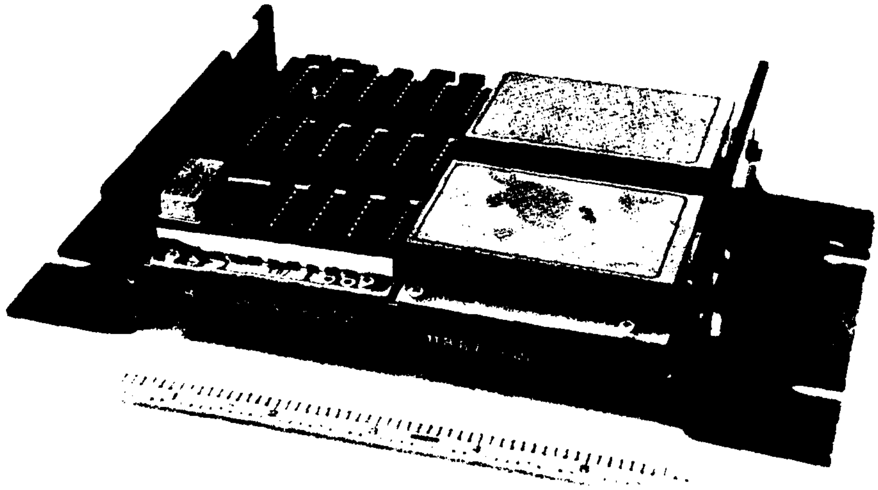


Figure 3-9. Standard Single Frame EECO Package

3.2.4 POWER SUPPLIES/CHASSIS DESCRIPTION

All DC power required by the fiber-optic transmit/receive unit (FOTRU) was supplied from modular power supplies mounted in a single, shielded enclosure directly below the standard EECO frame. This unit converts 115-V, 60-Hz AC power to +5 V DC, -5 V DC, and +12 V DC as required by the unit. The DC power is fed through the power supply enclosure via a filter network. Figure 3-10 shows the power supply, its shield cover, and the FOTRU circuits mounting frame. The entire assembly is then fastened together by screws, and a cover is mounted over the FOTRU frame.

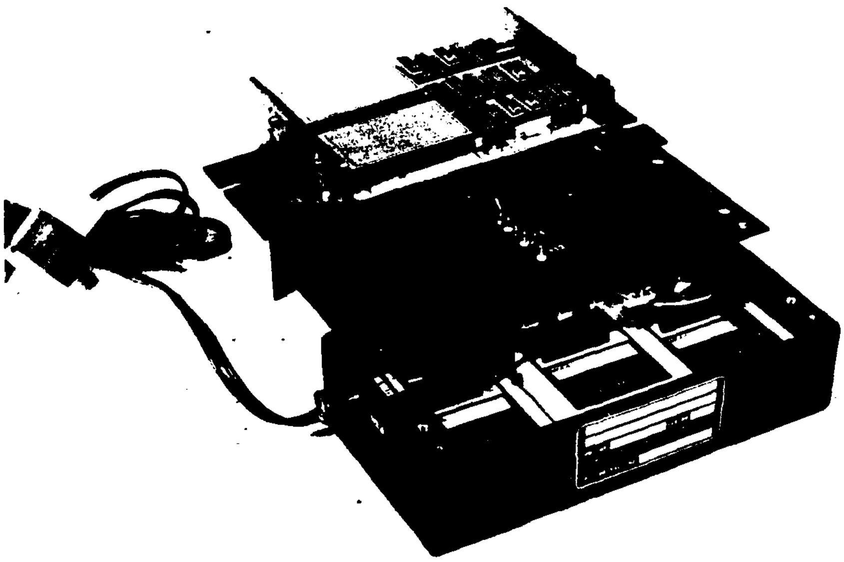


Figure 3-10. Power Supplies and FOTRU Circuits

The FOTRU is a total package containing all the circuitry and power supplies necessary to convert a 1553 Manchester-encoded electrical waveform into an optical waveform for transmission on the optical data bus. The bus is simply a multiport coupler connected by fiber-optic cables. Figure 3-11 shows two FOTRUs

connected in an optical bus configuration. Sixteen FORTRUs can be connected simultaneously to the multiport coupler.

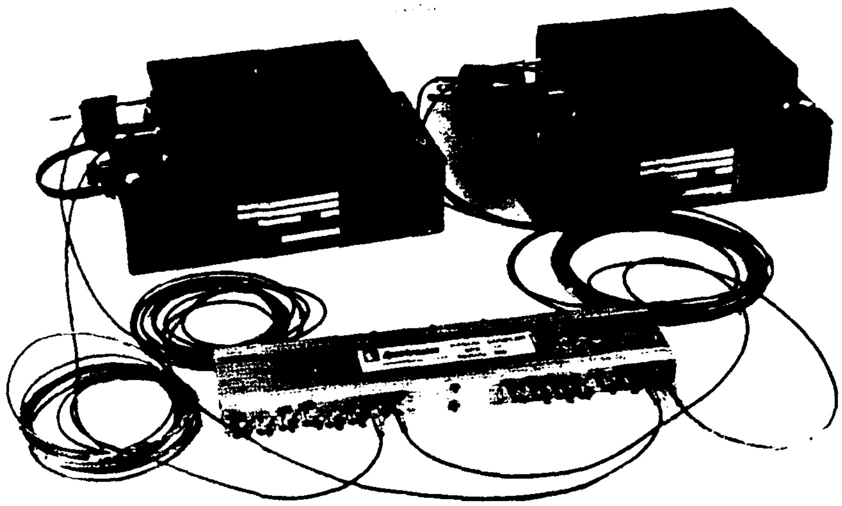


Figure 3-11. FOTRUs in an Optical Bus Configuration

3.2.5 SIXTEEN-PORT SYSTEM DESCRIPTION

Figure 3-12 shows all the hardware used in the systems testing of a 16-terminal 1553 Optical Data Bus in the A-7 Avionics Integration Laboratory. In this test, actual and simulated avionic hardware were interfaced on the bus. After completion of system testing, this hardware was delivered to the Naval Ocean Systems Center (NOSC) San Diego, CA.

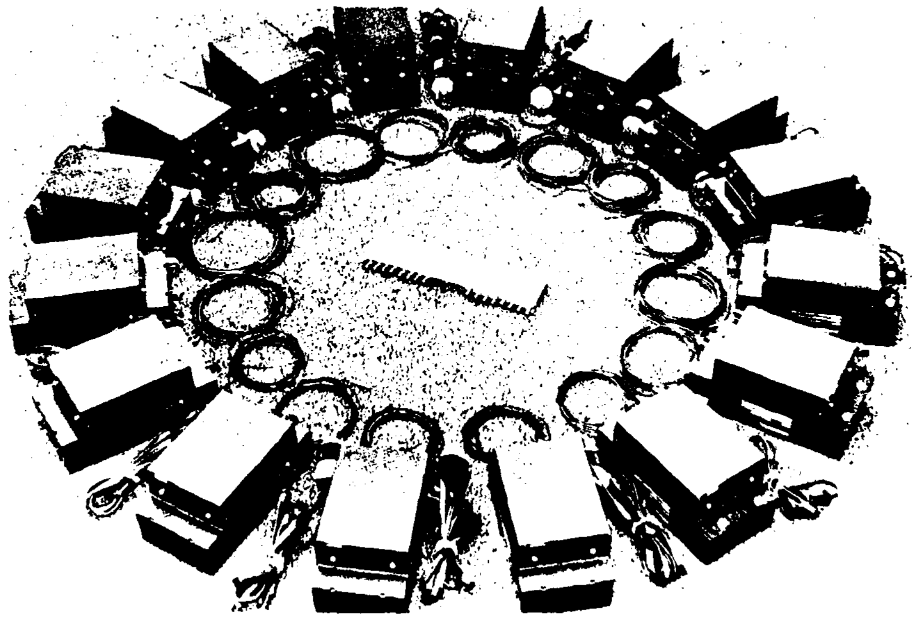


Figure 3-12. Hardware for 16-Terminal 1553 Optical Data Bus

3.3 TASK IA TEST CONFIGURATIONS

IBM's System Integration Facility (SIF) laboratory was the basic vehicle used to demonstrate the fiber-optic 1553A/B compatible data bus. Figure 3-13 is a block diagram of the SIF laboratory.

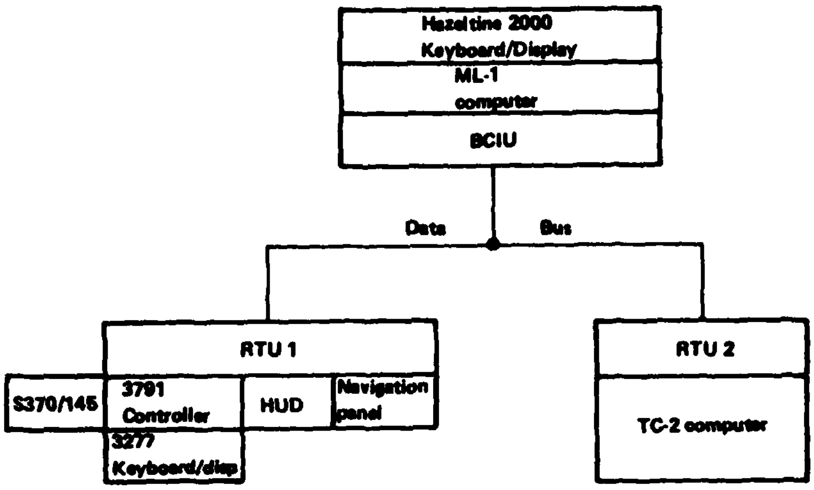


Figure 3-13. SIF Laboratory Block Diagram

The IBM ML-1 computer functioned as the bus controller, while the IBM TC-2 computer acted as the avionic onboard computer. Figure 3-14 shows the A-7 Bomb/Nav/Weapon Delivery System that was simulated in the SIF laboratory.

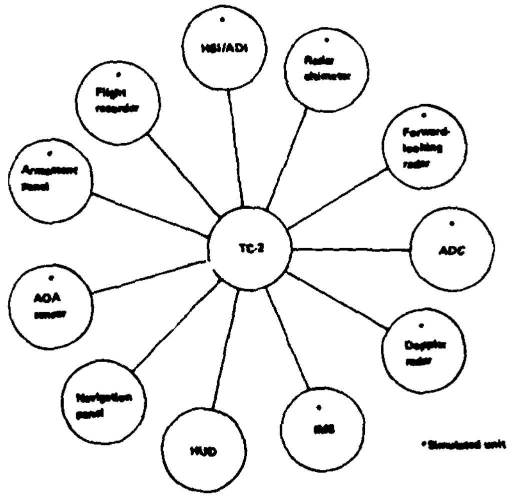


Figure 3-14. A-7 Bomb/Nav/Weapon Delivery System Block Diagram

The asterisked units were simulated in the S370/145 through the IBM 3791. Table 3-6 lists the data bus traffic, along with data words and messages per second.

Table 3-6. Data Bus Traffic

Type	Source	Sink	Messages per Second	Data Words per Second
RT-TO-RT	1	2	65	640
RT-TO-RT	2	1	89	1428
CT-TO-RT	ML-1	1 OR 2	15	240
RT-TO-CT	1 OR 2	ML-1	21	352

The bus controller interface unit (BCIU) and the two remote terminal units (RTU 1 and RTU 2) are normally interconnected through two redundant, standard, twisted-pair wire buses. These units were modified to replace one of the wire buses with a fiber-optic interconnection, as shown in Figure 3-15. One wire bus was kept to enable a switchover from the fiber-optic system to the electrical system (and vice versa in case of emergency).

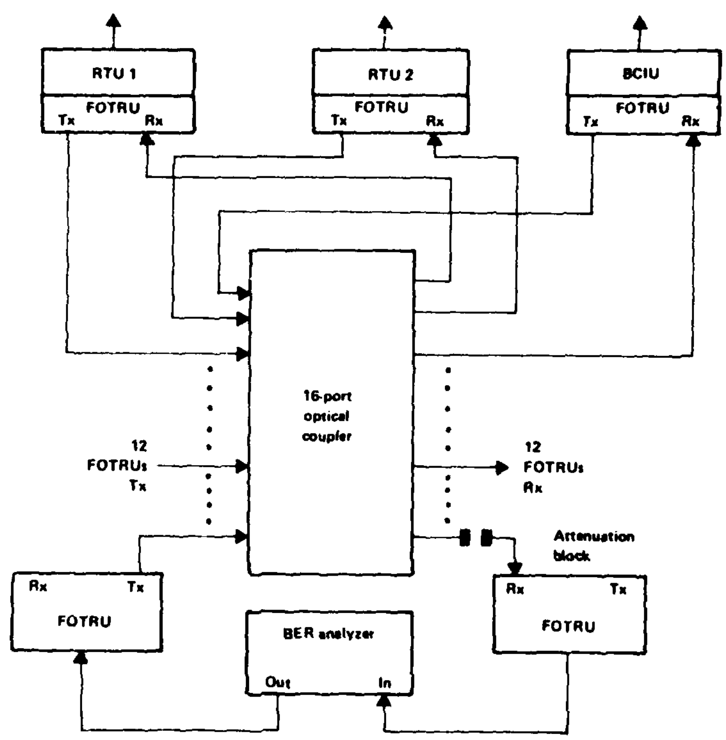


Figure 3-15. System Modification

RTU 1 interfaces with an IBM 3791 Processor; RTU 2 interfaces with an IBM 3277 Display Station, a military navigational computer, and a head-up display (HUD). The BCIU interfaces with a Hazeltine display, an ML-1 computer, and an ML-1 display through which software to operate the system is loaded. With the software, a simulated bombing run can be attempted, which involves communication between all displays and computers through the BCIU and RTUs.

The required software was first run in the system using the wire bus to confirm its correctness and to establish proper functioning of the system. The same software was then used in the same system, but now the FOTRU and optical cables replaced the wire bus.

In addition to the A-7 system, the remaining 13 ports on the fiber-optic bus were populated as shown in Figure 3-15. Twelve ports had inactive FOTRUs with power on, and the 13th port interconnected two FOTRUs in a configuration to measure the BER of the fiber-optic system.

3.4 TASK Ia TEST RESULTS

The Development and Test Plan of MIL-STD-1553A Compatible Fiber-Optic System, Revision B, dated 2 March 1979, describes the various tests performed during the Task Ia hardware demonstration. Certain tests, particularly those involving the recommended 19-fiber cable, could not be performed because of the lack of a cable termination technique. Instead of the 19-fiber cable, some Gallite 2000 (500 dB/km) was purchased in 3-m lengths and used for component, subsystem, and system level testing. Those sections of the test plan that were not implemented are noted in the test result descriptions that follow.

3.4.1 COMPONENT LEVEL TESTING

3.4.1.1 Light-Emitting Diodes

Table 3-7 lists the IBM-measured light output, along with the vendor results for each of 20 SE2231 LEDs. Vendor measurements were taken through a physical aperture of 1.14 mm and into a numerical aperture (NA) of 0.24, whereas IBM measured total light output.

Table 3-7. Measured Light Output

LED	Power Output (mW) IBM	Vendor	Transmitter Serial No.
A	3.69	1.11	T017
B	3.15	0.90	T012
C	3.06	0.78	T019
D	2.40	0.71	T007
E	3.33	0.94	T010
F	3.24	0.95	T001
G	2.94	0.72	T004
H	3.60	1.05	T006
I	3.42	0.98	T013
J	3.33	0.90	T014
K	2.85	0.81	T016
L	3.64	0.94	T005
M	3.44	0.89	T009
N	3.28	0.86	T018
P	3.19	0.90	T002
R	2.94	0.90	T011
S	3.64	0.90	T003
T	3.24	1.06	T015
U	3.24	1.02	T008
V	2.52	0.71	Spare

3.4.1.2 Input Coupling

The test plan called for a mode stripped 19-fiber bundle cable. As previously stated in Section 3.2, Gallite 2000 was used instead. The attenuation of the short cable was accounted for in the results.

Because all the FETs used in the original assembly were defective and had been replaced in the hybrid transmitter, the actual input coupling losses are in doubt, but cannot be measured because the hybrids have been completely assembled. The average input coupling loss as measured before was 6.9 dB. This number will be assumed for the remaining transmitter measurements.

3.4.1.3 Output Coupling

To determine the output coupling (fiber-to-photodiode), we assumed the published responsivity for the photodiode (0.5 A/W). With a known light power (DC) out of a cable (measured with the integrating cube) connected to each receiver, the change in voltage across a

known resistance in series with the photodiode was measured and the diode current for the known light input calculated. The following formula was used to calculate the input coupling loss in dB:

$$L = 10 \log \frac{P_r \times r \times R_s}{E \times 10^6}$$

where P_r is the known light power in mW

r is responsivity (0.5)

R_s is the series resistance in ohms

E is the change in voltage in V

Measurement of 16 receivers gave an average output coupling loss of 3.2 dB, with values ranging from 2.7 to 3.9 dB.

3.4.1.4 Cable Measurement

All of the cables used (Gallite 2000) were 3 m long. The attenuation of 500 dB/km was verified, generally, by measurement of several cables. Because even for 3-m lengths the fiber attenuation could not be neglected (≈1.5 dB), and because at least one connection loss would be included when using the substitution method, the loss caused by the cable and the connection could not be separated. Several 3-m cables with one connection gave measured losses of nominally 5 dB. With possible connector losses of 3 to 4 dB for this type cable, this measurement was reasonable.

3.4.1.5 Connector Measurement

As just stated, all of the cables were 3 m long with some attenuation. Because attenuation of the connection and the attenuation of the fibers is not separable, an attenuation of 3.5 dB was assumed for the connection (a value which is consistent with previous industry experience for this type cable) and 1.5 dB for the cable (500 dB/km).

3.4.1.6 Multiport Coupler

Again, by substitution, all possible combinations of input and output port attenuations were measured (256 in all). Figure 3-16 is a histogram of the measurements, the greatest number of which (86) gave a loss (including input and output connection) of 20.3 dB, with a range from 19.3 dB to 22.1 dB. If the connector loss of Paragraph 2.1.5 is subtracted, the coupler loss for the most likely value is (20.3 - 3.5) 16.8 dB, where 12 dB is the loss caused by 16-way power division, and 4.8 dB is the excess loss.

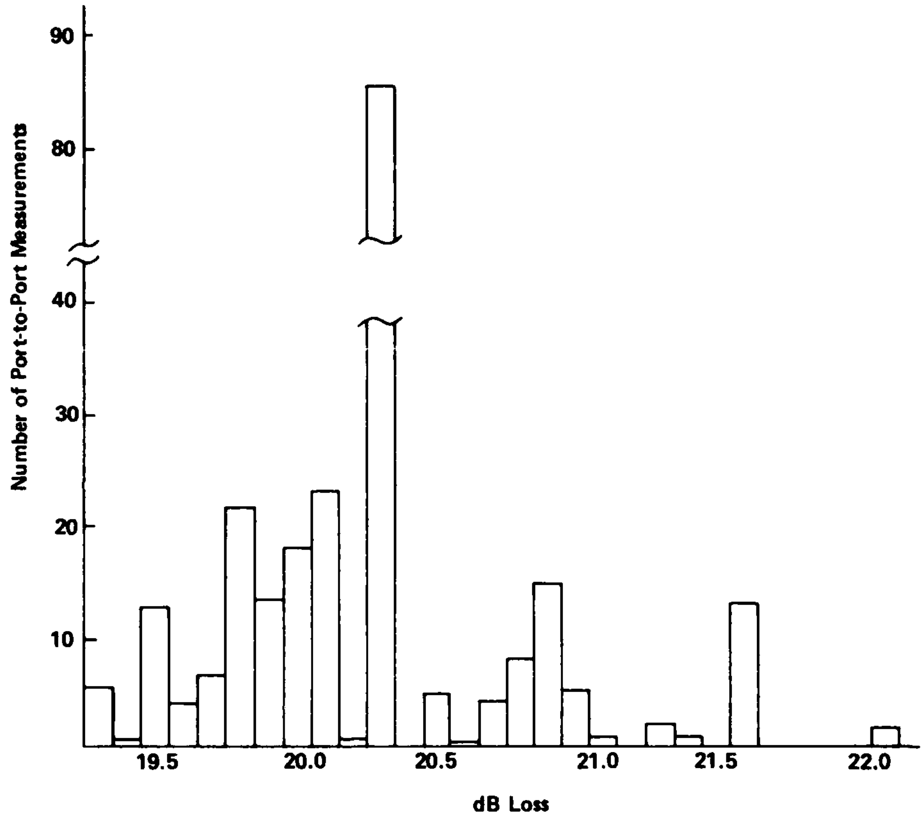


Figure 3-16. Multiport Coupler Histogram

3.4.2 SUBSYSTEM LEVEL TESTING

3.4.2.1 Transmitter

Each transmitter was measured with a calibrated receiver to determine the peak pulsed light output. It was assumed that the input (LED to cable) coupling loss was 6.9 dB (Paragraph 3.4.1.2). A known attenuation was inserted between the transmitter and the calibrated receiver to stay within the linear region of the preamplifier gain. The voltage at the output of the

transimpedance preamplifier was measured, and the following formula was used to determine the peak pulse light output from the LED:

$$P_o = \frac{(\Delta E_T) \times 10^{\left(\frac{L_I + L_A + L_o}{10}\right)}}{G_T \times r} \qquad \text{(mW)}$$

where L_I = input coupling loss in dB (6.9)

L_A = known attenuation in dB (17.1)

L_o = output coupling loss in dB (2.9)*

r = photodiode responsivity (0.5)

G_T = transimpedance amplifier gain (4,300)

ΔE_T = transimpedance amplifier output in mV

*Terminal serial number 2.

Table 3-8 gives the transmitter peak power outputs for a 1-A, 50-ns current pulse input to the LED. The nominal rise time was 7.5 ns, and the transimpedance amplifier voltage output rise time was 12 ns. The bandwidth of the preamplifier is approximately 20 kHz to over 60 MHz at the -3-dB gain points. Because of this, the output rise time is a very good representation of the light rise time of the LED.

Table 3-8. Transmitter Peak Power Outputs

Transmitter Serial Number	LED Peak Output (mW)
T001	15.9
T002	15.9
T003	13.2
T004	10.3
T005	13.7
T006	13.2
T009	12.5
T011	15.9
T012	15.9
T013	15.3
T014	12.1
T015	16.7
T016	13.7
T017	11.4
T018	13.7
T019	9.8

3.4.2.2 Receiver

3.4.2.2.1 Preamplifier

The low- and mid-frequency gain of the transimpedance preamplifier is a function of the collector resistor, R_c, the feedback resistor, R_f, and the transistor forward current gain, β. Figure 3-17 is a schematic of the receiver preamplifier.

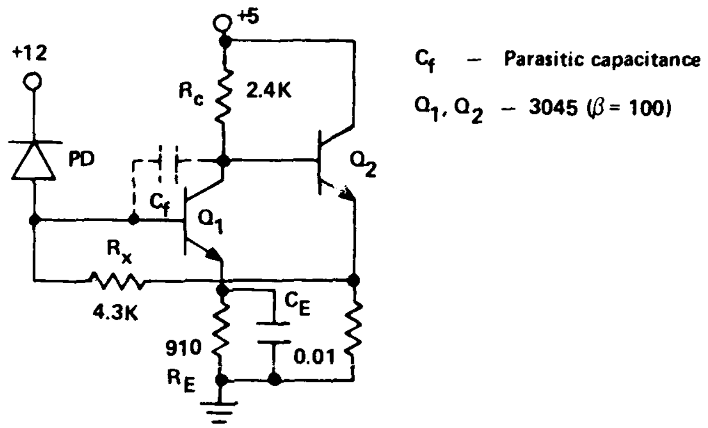


Figure 3-17. Receiver Preamplifier Schematic

At frequencies, f, where

$$\frac{1}{2\ R_E C_E} << f << \frac{1}{2\ \pi\ R_f C_f},$$

$$\frac{e_{out}}{i_{in}} = \frac{\beta R_c R_f}{\beta R_c + R_f} = \frac{R_f}{1 + \frac{R_f}{\beta R_c}}$$

In this case, $\beta R_c = 100 \times 2400 = 0.24 \times 10^6$

Because $\frac{R_f}{\beta R_c} = \frac{0.0043 \times 10^6}{0.24 \times 10^6} = 0.18 << 1$,

$$\frac{e_{out}}{i_{in}} \cong 4.3\ K\ \Omega$$

Because $i_{in} = rP_i$ where r = diode responsivity (0.5)
P_i = light power in mW

$e_{out} = 4.3 \times 10^3 \times 0.5 \times P_i$ (mW)

3.4.2.2.2 Postamplifier

The gain of the postamplifier was measured using the method shown in Figure 3-18.

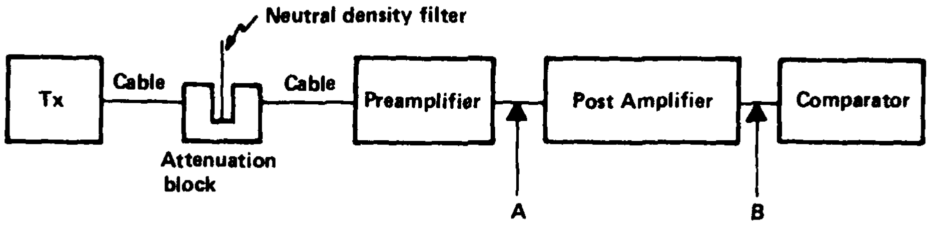


Figure 3-18. Postamplifier Gain Measurements Block Diagram

With no filters in the attenuation block, the peak voltage at point A in Figure 3-18 was adjusted to the upper end of the linear region of the preamplifier by adjusting the spacing between the two cable ends in the attenuation block while the transmitter (Tx) was operating in a normal mode with 50-ns pulses once every microsecond. The peak voltages at A and B were recorded. Known steps of attenuation were inserted into the attenuation block and the peak voltage at B recorded at each step. This method was necessary since, at low light levels, the voltage at A is too small to measure with normal laboratory oscilloscopes.

A gain curve was then plotted of B as a function of A. Figure 3-19 shows the composite plots for all 16 receivers for the linear region. The average gain of the postamplifier was 234 in terms of pulse peak voltage. Figure 24, page 93 of the Task I Analysis Review Report shows (for the dotted curve) the actual peak pulse amplitude to be 0.6 of the value of the single frequency gain between the -3-dB points of the receiver bandwidth. Therefore, the real single-frequency gain of the receiver is 234 ÷ 0.6 or 390.

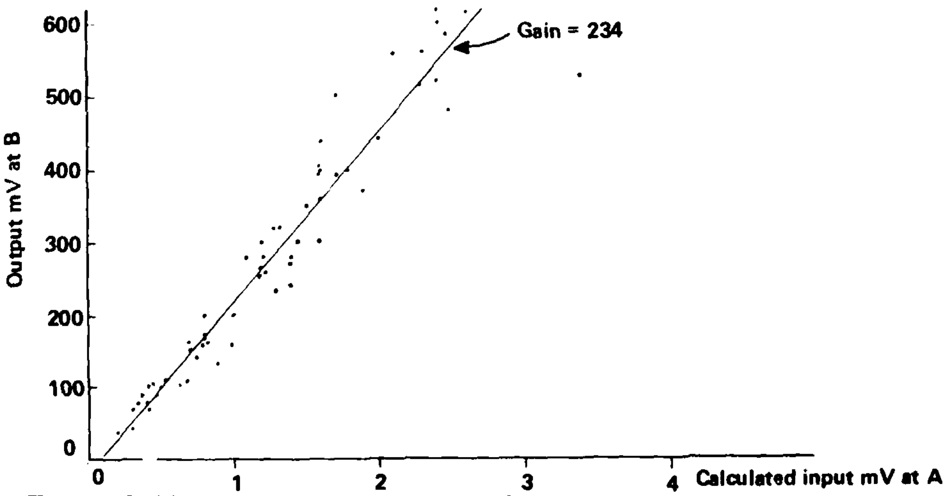


Figure 3-19. Composite Receiver Plots

3.4.2.3 Receiver Noise

Measurement of noise was impossible at the output of the preamplifier because it was at too low a level. The measurement was therefore made at point B or the input to the comparator. The method used is described in "Don't Eyeball Noise," by G. Franklin and T. Hatley, Analog Systems, Data Disc, Inc., published in the November 22, 1973 issue of Electronic Design. The measurements of noise varied from 7 mV to 10 mV rms at point B.

Using the formula for noise developed in Appendix B for a bipolar transimpedance amplifier and the gain characteristics of the post-amplifier, one calculates the noise at point B to be 9.9 mV rms. This correlates rather well with the measurements.

3.4.2.4 Receiver Sensitivity

Receiver sensitivity is defined as the light power level in dBm at the input which will produce a signal equal in amplitude to the rms noise at the output point of measurement (point B).

From Sections 3.4.2.2.1 and 3.4.2.2.2, the total receiver pulse gain can be written as

$$\frac{e_{out}}{P_i} = 234 \times 0.5 \times 4300$$

Therefore,

$$\text{receiver sensitivity} = 10 \log \frac{P_{i1}}{10^{-3}}$$

where P_{i1} is power in for a signal-to-noise ratio of one

Because $P_{i1} = \dfrac{0.0085}{234 \times 0.5 \times 4300} = 1.69 \times 10^{-8}$ W, (0.0085 V is average noise)

receiver sensitivity = -47.7 dBm

3.4.2.5 Receiver Summary

The receiver characteristics are summarized in Table 3-9.

Table 3-9. Receiver Characteristics

Photodiode responsivity	0.5 A/W
Preamplifier transimpedance	4300 Ω
Postamplifier gain	
Peak pulse gain	234
Single midfrequency gain	390
Total receiver responsivity	
Pulse	5×10^5 V/W
Single frequency	8.4×10^5 V/W
Total receiver responsivity (including cable output to photodiode coupling loss	
Pulse	2.4×10^5 V/W
Single frequency	4.0×10^5 V/W
Receiver sensitivity	-47.7 dBm

3.4.3 LINK MEASUREMENTS

3.4.3.1 Link Delay

The nominal link delay from Manchester in to Manchester out with 6 m of cable and one coupler is 500 ns. Delay allocation is

Component	Delay (ns)
Transmitter logic	190
Transmitter analog	10
Cable and coupler	30
Receiver analog	40
Receiver logic	230
Total	500

3.4.3.2 Pulse Distortion and Jitter

With the signal at the input to the comparator in the receiver set at the lowest level for 10^{-12} BER, the pulse leading-edge jitter at the Manchester output was less than 5 ns.

3.4.3.3 Bit Error Rate (BER)

Figure 3-15 shows the BER test setup. The block labelled Bit Error Rate Analyzer was a Hewlett Packard Unit. To obtain consistent readings and to be able to extrapolate signal requirements of 10^{-12} BER, equally spaced pulses at 1 ns, with duration of 50 ns, were input to the transmitter. This represents the average pulse spacing for any Manchester word or message. The optical signal to the receiver was degraded with filters as shown until a measurable BER occurred.

From error function (erfc) tables, the signal-to-noise ratio (SNR) rms required for that BER was taken. Extrapolation to 10^{-12} BER was then made by finding the signal level to increase the SNR to that required, comparing it to the actual signal for the measured BER, and subtracting that ratio in terms of dB from the actual attenuation for the measured BER.

In summary,

- With 46.2 dB attenuation, BER was 2.2×10^{-5}.
- BER of 2.2×10^{-5} requires SNR of 4.08.
- BER of 1×10^{-12} requires SNR of 7.03.
- *With 45.5 dB attenuation, BER will be 1×10^{-12}.

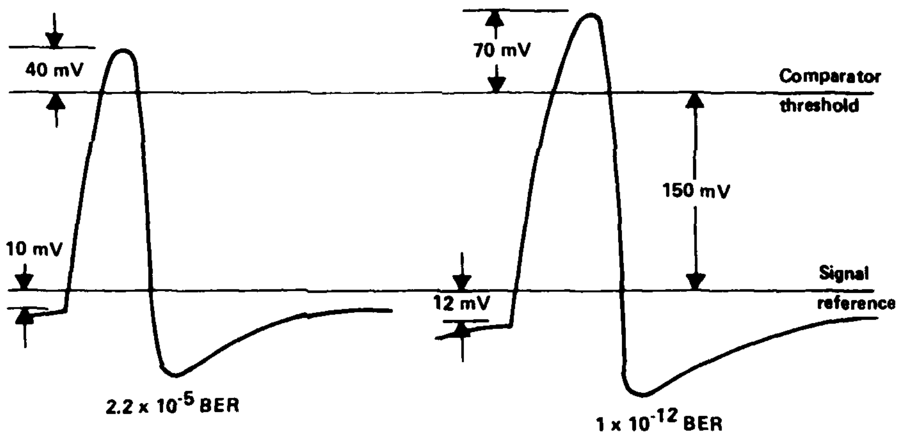


Figure 3-20. BER Waveform

* Figure 3-20 shows the waveforms and requirements for the two BERs. The SNR referred to above is the signal above threshold, to rms noise ratio. Worst-case noise at this point is 10 mV rms. A BER of 2.2×10^{-5} requires a SNR of 4, or 40 mV above threshold. A BER of 1×10^{-12} requires a SNR of 7, or 70 mV above threshold.

As depicted in Figure 3-20, to obtain the total signal amplitude, the 150-mV theshold must be added to that portion above the threshold, and that portion of the AC-coupled signal below the signal reference must also be added. The total peak signal for 2.2×10^{-5} BER is 200 mV and that for 1×10^{-12} BER is 232 mV. This is a ratio of 1.16:1, or 0.7 dB. To achieve a 1×10^{-12} BER would require no more than (46.2-0.7) 45.5 dB attenuation.

3.4.4 SYSTEM LEVEL TESTS

During the checkout of the MIL-STD-1553 fiber-optic data bus in the IBM System Integration Facility, a simulated Navy A-7 aircraft bombing run was demonstrated using both the electrical and the optical buses. The simulation was successful in both cases. Based on at least this test, the fiber-optic system can be described as equivalent to the electrical system. Since no bus-related error was detected on either bus, it is not possible to attribute superior performance to either system based on this test.

Figure 3-21 shows the logical Manchester data word input, the plus threshold, zero crossing, and minus threshold logic outputs, in that order, from the optical system. Figure 3-22 compares the logical outputs from the electrical and optical buses, with the center signal being the zero crossing signal. Obviously, at this point in the system, there is no difference between the two.

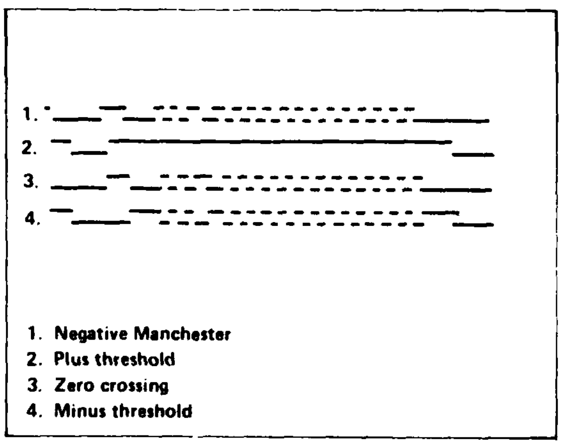


Figure 3-21. Manchester Data Word Input

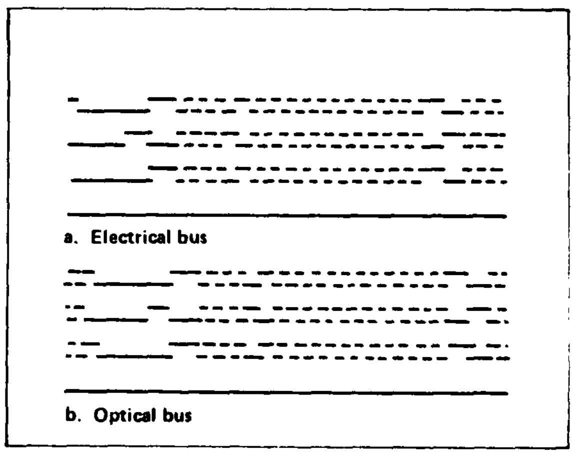


Figure 3-22. Electrical and Optical Bus Output Comparison

A bit error rate test was conducted with the optical system. This is described in Paragraph 3.4.3.3 of this report. A bit error rate (BER) of 1 x 10^{-12} was proven by extrapolation. A similar demonstration on the electrical bus was not possible because there is not an equivalent way to insert a known electrical attenuation and know quantitatively or precisely what one has done to the signal on the bus.

A way of comparing the two systems is to examine the 1553B requirements for the electrical bus and compare them to the actual operating characteristics of the optical bus. It will be shown that the optical bus will always provide conditions which result in a lower bit error rate with the same logical input to the transmitter as in the electrical system.

Paragraph 4.5.2.2.1.2 on page 29 of MIL-STD-1553B defines requirements for the electrical output waveform from a terminal to the bus. This would be the optimum input to a terminal if one assumes no degradation on the bus. Paragraph 4.5.2.2.2.4 requires a 1 x 10^{-7} word error rate in the presence of 200 mV rms noise.

Figure 3-23 is an idealized waveform with the worst-case conditions as described in paragraph 4.5.2.2.1.2 and 4.5.2.2.2.4 of the standard for meeting a word error rate of 1 x 10^{-7}. The peak-to-peak threshold SNR is 7.5, and allowable zero-crossing-to-zero-crossing tolerance is ±25 ns.

During the detection process, the receiver clock is synchronized to the zero-crossing time of the data or command sync. Under the worst-case conditions shown, the sync zero crossing will be at A ±175 ns with a probability of 1 - 0.3 x 10^{-13} (1-erfc 7.5), and the first data bit zero crossing will be at B ±150 ns.

If the receiver clock is "perfect" relative to the incoming data, the clock sample times will be 500 ns apart. Because the clock is "synchronized" by the data sync zero crossing, points A and C (the center of the first half-bit or ideal sample time) will differ in time by 1750 ±175 ns with the same probability (1 - 0.3 x 10^{-13}). This is a 350-ns ambiguity and will therefore cause the data bit to be in error with a probability much greater than (0.3 x 10^{-13}) because, as shown in Figure 3-23, the width of the first half-bit can be as narrow as 200 ns.

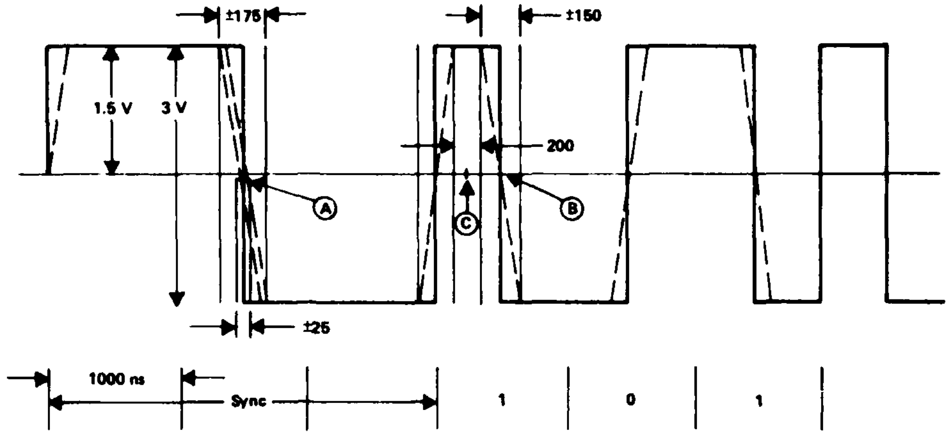


Figure 3-23. Worst-Case Condition Idealized Waveform

To have an error probability compatible with all signal tolerances, these tolerances cannot cause a time difference ambiguity between points Ⓐ and Ⓒ of greater than the narrowest half-bit width with a probability greater than that required for the chosen BER. Either the risetime specification (300 ns), or the jitter specification (±25 ns), or the peak signal-to-noise specification, or all could be tightened to achieve the BER implied by the given SNR.

The other alternative (what is actually done) is to design the receiver to limit the noise. To determine the BER for the system as specified, one can go down from the peak signal level to the point where the sync zero-crossing ambiguity is exactly equal to the minimum first half-bit width, compute the peak signal-to-threshold-noise ratio at that point, and look it up in the erfc { x } tables, where { x } is that SNR. In the case shown (Figure 3-24), it turns out to be at the 3/4 amplitude level where the minimum first half data pulse width is 275 ns centered at Ⓒ , and the sync zero crossing is ambiguous by ±137.5 ns relative to point Ⓒ . This SNR is 5.625 (1.125 V ÷ 0.2 V).

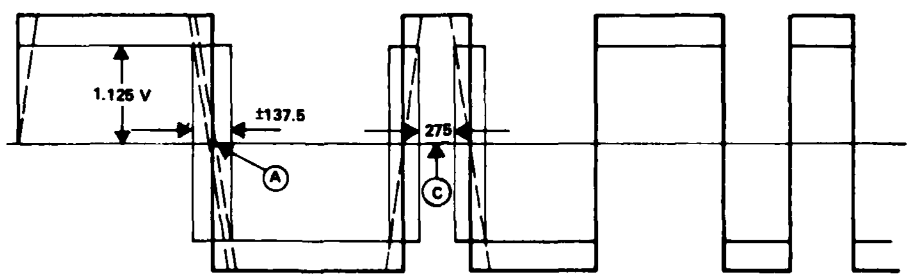


Figure 3-24. Actual Waveform

The BER under these conditions is 9.37×10^{-9}. Because for a 20-bit word, a BER of 5×10^{-9} is required, for a word error rate (WER) of 1×10^{-7}, the noise must be reduced so that at the 1.125-V signal-to-threshold level, the SNR is 5.74. The noise should be $1.125 \text{ V} \div 5.74 = 196 \text{ mV}$. The noise bandwidth specified in 1553B for the 200-mV rms noise is 1 kHz to 4 MHz.

To reduce the noise to 196 mV requires a bandwidth of 3.8 MHz. A low-pass filter with an upper cutoff frequency of 3.8 MHz can accomplish the required results. Achieving higher BERs would require narrower bandpass filters, which also affect the signal itself. A well-designed receiver can accomplish the 10^{-7} WER, but not much better under the restriction of the 1553B electrical waveform specifications as interpreted herein. It would be even more difficult, if not impossible, to achieve the specified WER if one assumed Paragraph 4.5.2.2.2.1 of MIL-STD-1553B as the worst-case input.

Figure 3-25 shows the waveform limits for the IBM pulse-position-encoded optical Manchester data as it appears on the optical bus. The dotted waveform is the 1553B STD used as a reference. The measured leading edge jitter of the decoded Manchester in the IBM receiver was less than 5 ns when the optical signal was set at the minimum amplitude for 10^{-12} BER. In the specification for the IBM transmitter, the rise time of the optical pulse will be less than 20 ns and the leading edge jitter will be ±7.5 ns.

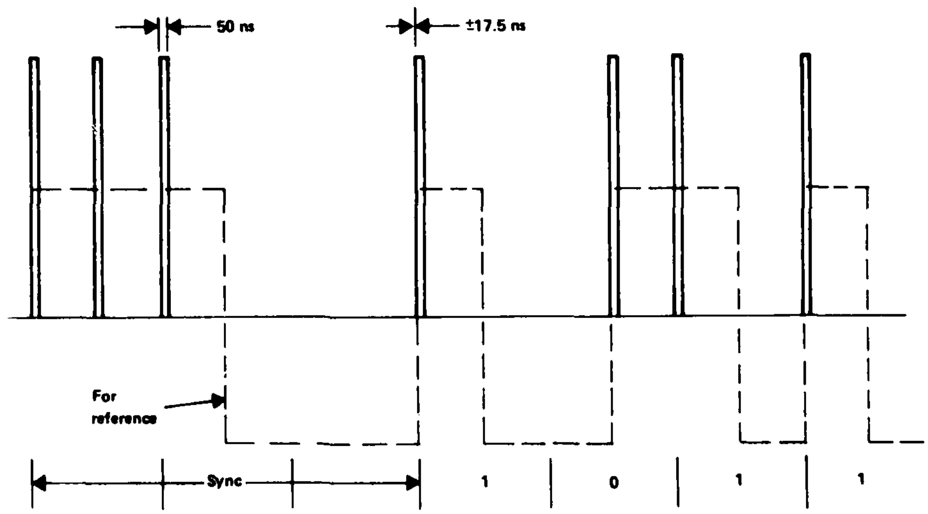


Figure 3-25. IBM PPM Manchester Data Waveform Limits

Digitally, in the receiver, the original Manchester bilevel signal is reconstructed by triggering a comparator above some threshold with the amplified electrically converted optical signal, and setting a logical single-shot for 500 ns. This single-shot is actually a counter which is started by the comparator output gating a 20-MHz clock and stopped when the count reaches 11. Its accuracy is a function of the 20-MHz oscillator stability. For all practical purposes, the length of the reconstructed Manchester pulse is exactly 500 ns. For a given SNR at the comparator input, because the pulse rise time can be 20 ns and its location in time ambiguous by ±7.5 ns, the location in time of the 500-ns pulse relative to a reference can be ±17.5 ns with a probability consistent with that of the SNR. In a similar manner, the reconstructed sync can have a zero-crossing ambiguity of ±17.5 ns.

Within this condition and for whatever BER results from the given SNR, the Manchester bit sampling time can only be ±35 ns from the center of the half-bit time. Because the reconstructed pulse is 500 ns wide, the sample time will always be well within the pulse envelope. What is significant in the IBM approach is that the BER is a function of the SNR only and not the jitter or risetime of the optical pulse.

Another advantage of the optical bus is that the noise at the receiver is a function of the receiver design and under control of the designer to the extent that he or she can minimize it for a given system and knows what it is. Noise sources outside the terminal cannot affect the optical bus as they can the electrical bus.

With the IBM PPM Manchester optical approach, BERs of 10^{-12} are practically achievable when system specifications require such error rates, and when it is also required to be compatible with all but the transmission media of MIL-STD-1553B.

Section 4
TASK II DETAILS

This section describes the studies leading to the "typical" 1985-1990 avionic system, which in turn produced a set of requirements for system interconnection. This section also describes the tradeoffs made that resulted in a recommended implementation of a fiber-optic, 50-Mb serial data bus.

4.1 INITIAL ASSUMPTIONS

Two basic groundrules were agreed to by the Navy and IBM as follows:

- This study was to be oriented toward avionic systems only.
- If technology predicted to be available as production items in 1985-1990 were unavailable in any form today, it could be simulated or the system could be operated at some lesser degree of performance so long as feasibility of the 1985-1990 system were proven.

4.2 TASK II METHODOLOGY

Figure 4-1 is a flow diagram illustrating the approach to the Task II study.

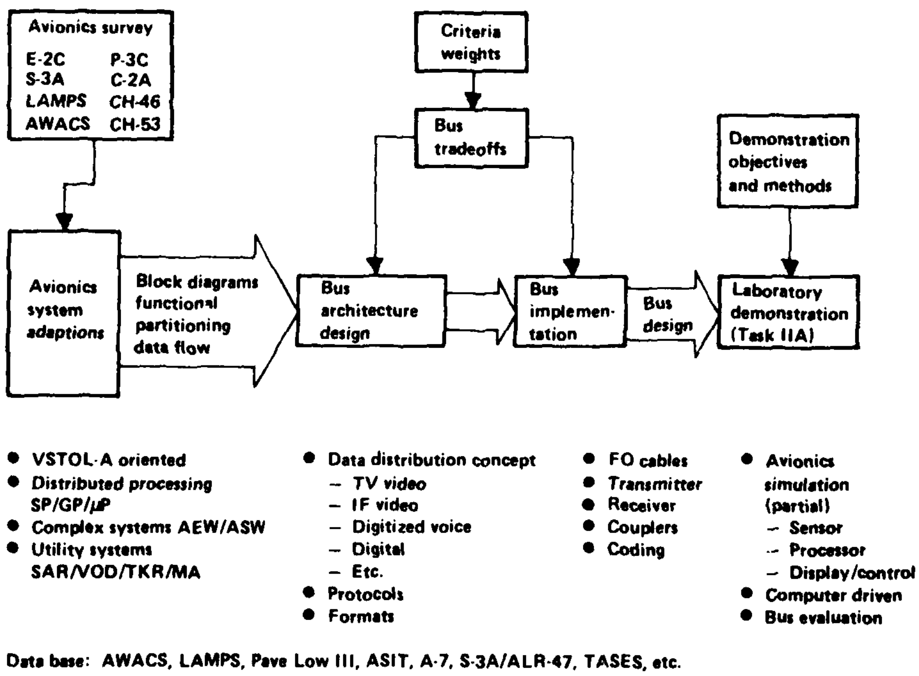


Figure 4-1. Task II Methodology

Based on IBM's advanced systems studies oriented toward VSTOL-A, and on previous and on-going avionic contracts such as AWACS, LAMPS, S-3A, E-2C, etc., a "typical" 1985-1990 IOC Navy avionic system was synthesized and partitioned for data flow. Interconnect architectures and fiber-optic implementations were traded off based on projected technologies and data distribution concepts. Finally, a recommended fiber-optic interconnect system (bus protocol and architecture) was presented to the Navy for their approval. Concurrently, a laboratory demonstration and test plan (contract data item A009) was presented for the optional Task IIa.

4.3 TYPICAL SYSTEM DESCRIPTION

Digital interconnect structures (buses) represent a major and sometimes limiting factor in the data-handling and control capability of any processing system. Over the last 10 years, the data processing portion of an avionic system has grown from a central bombing/navigation computer to a sophisticated array of data processors, handling various types of sensor data. Traditional analog methods of signal filtering and analysis have been replaced by high-speed digital array processors. The modern avionic system, with its digital signal processing subsystem, requires several levels of digital interconnect structures to optimize the flow of data and control to the various system elements.

This section presents a "typical" 1990 avionic system. Although this system may never exist, it illustrates the mix of processing elements that a future avionic system might experience. The typical system is analyzed, and data-handling requirements are derived. Particular attention is paid to the command and control flow between the various system elements. The analysis of the avionic system is divided into two subsections: a signal processing subsection, and a general-purpose data processing subsection.

Any analysis that depends on future projections for its base remains somewhat subjective. To provide some solid check points, the bulk data bus requirements are derived by extrapolating two current processors: the AN/UYS-1 Advanced Signal Processor, and the NATO/E-3A, Model CC-2, General-Purpose Processor.

4.3.1 PROJECTED 1990 AVIONICS SYSTEM

Figure 4-2 illustrates the system block diagram for an advanced ASW aircraft such as VSTOL or MPA. The most important characteristic of this system is that the processing load is shared between many separate processors.

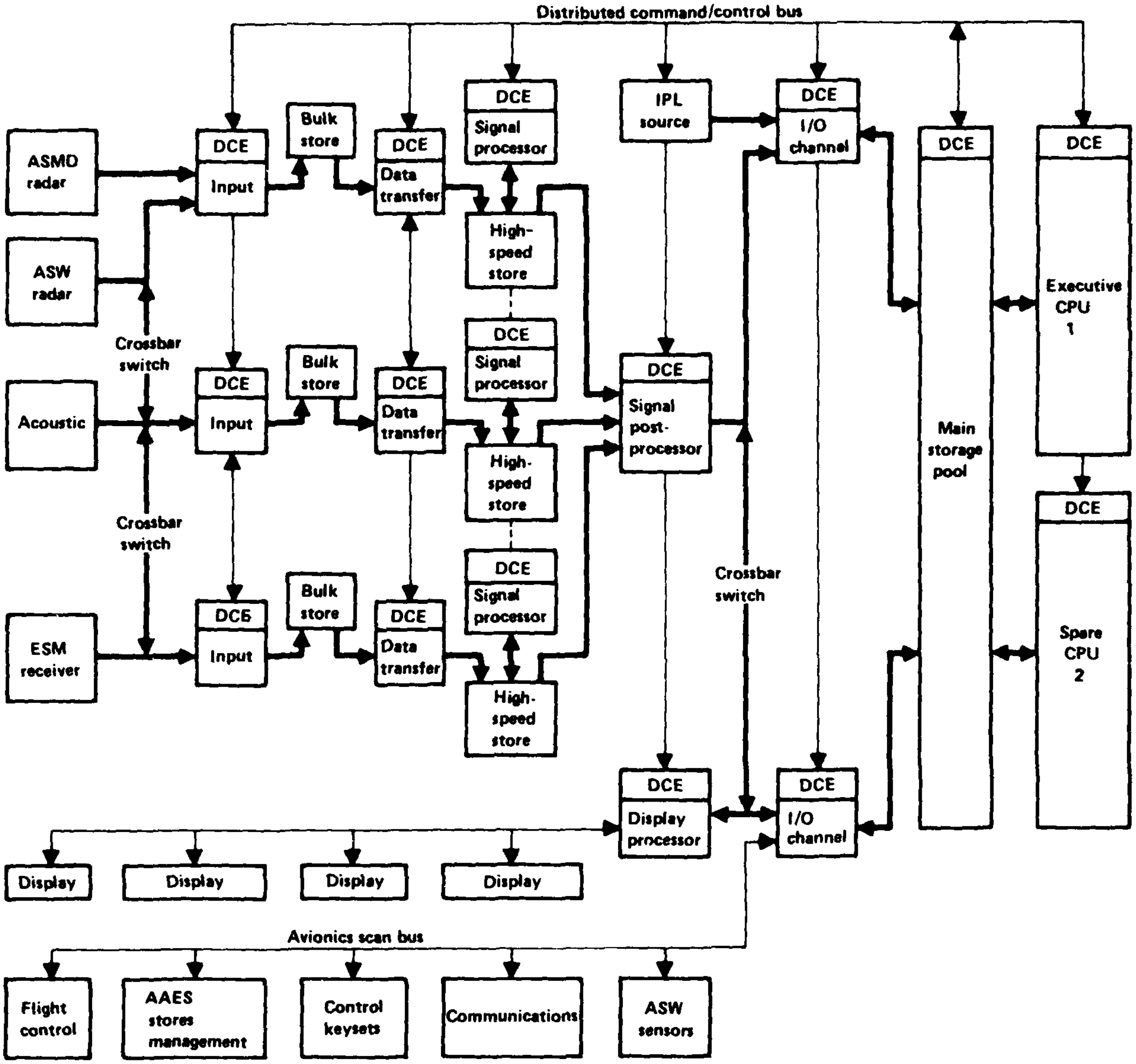


Figure 4-2. Projected ASW Avionic System (C 1990)

The system is distributed in the sense that, as the data flows in from the sensors toward the general-purpose (GP) processor, it passes through several different special-purpose processors. Each processor in the chain operates on and compresses the sensor data and passes it to the next processor. When the data reaches the GP processor, it has been reduced to a highly condensed format suitable for command/control decision making (i.e., target reports).

The system is centralized in the sense that all executive level decision making is contained in the GP processor. An important characteristic of the GP machine is its ability to branch quickly. The GP processor manages the system configuration by assembling command blocks that link together major subroutines contained in the special-purpose processor's own local storages. The system control/linkage is performed over the distributed command/control bus. The system command linkage as it relates to the distributed bus is detailed in the following sections.

System synchronization is distributed. Each processor communicates with adjacent processors in the chain via the distributed command control bus. When a processor has completed a major data frame, it signals to its downstream neighbor by transmitting a sync message on the distributed command/control bus. Processors key major operations on sync messages.

The GP processor is as loosely coupled to the other system elements as possible. Once a system mode has been established, the GP machine simply gathers in the results, but does not get involved in the fine detail of system synchronization and data management. The GP machine must, however, maintain overall control, and accomplishes this task by assembling higher level command blocks for major system modes.

Mission survivability is assured by the use of redundant "hot spare" system elements and redundant data paths. A spare GP processor can be configured in to maintain executive control should a GP failure occur. Crossbar switching of data paths is provided at the inputs to the three signal processors and the I/O channels. Should a signal processor or I/O channel fail, the remaining elements can be switched in to carry the failed element's processing function in a degraded mode.

The design of any large system of this type must consider data flow, control, synchronization, and survivability/redundancy. Although the system design approaches can vary, these factors must be addressed by any final design.

The data rates for the various system sensors that provide inputs to the avionic processing complex are tabulated in Table 4-1.

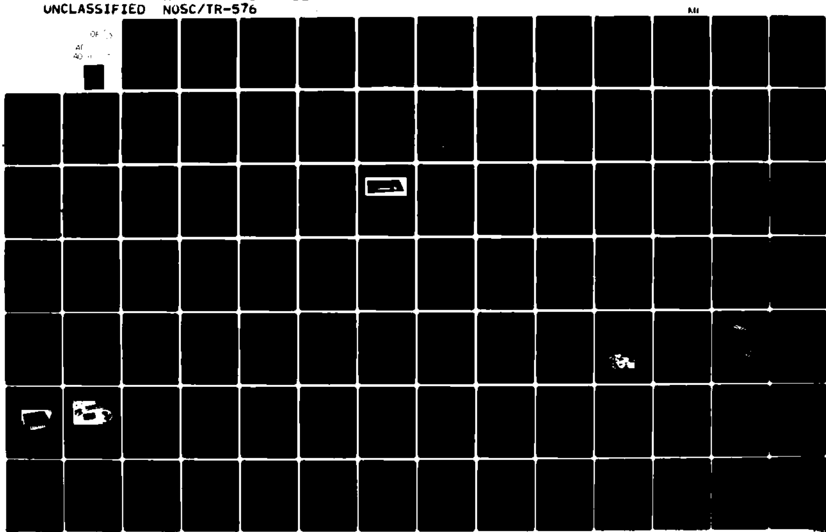
Table 4-1. Avionics Processor Data Rate Estimate
Sea Control/Open Ocean/1990-2000

Avionics	Input		Output		Avionics	Output		Input	
	kb	R	kb	R		kb	R	kb	R
Doppler	*	*	0.3	A					
Magnetic compass	*	*	0.3	C	Bulk store	10,000	1	10,000	1
Laser	*	*	0.3	C	Buffer store	320,000	1	320,000	1
IR	7.7	D-E	2.6	D	Memory	320,000	1	320,000	1
MAD	2	D	2	D					
Acoustic	*	*	10,000	1	SO	3	C	0.4	A
Radar (ASW)	0.8	B	3,000	1	ATO I	3	C	0.4	A
IFF Interrogator	*	*	-	-	ATO II	3	C	0.4	A
					Pilot	3	C	0.4	A
Actuators	6	B-E	6	B-E	Display controllers	*	*	80	C
Engine sensors	*	*	*	*					
Controls	*	*	2.5	D-E	VHF	16	C	16	C
Radar altimeter (3)	*	*	2	C	Security	48	C	48	C
Rate gyros (3)	*	*	20	D-E	ICS	32	C	48	C
Accelerometers (3)	*	*	20	D-E	ILS	0.3	B	*	*
Air data sensors (3)	*	*	4	C	AWCLS	0.8	B	*	*
Flight control computers (3)	12	D	120	D	UHF	48	C	32	C
					IFF transponder	*	*	*	*
AAES	0.5	1	0.5	1	JTIDS	16	1	16	1
Stores management	0.7	1	0.7	1	TACAN	1	C	*	*
					Radar (AEW)	650,000	1	15	E
					ESM	36,000	1	3	C

Refresh rate

A - 5/s D - 50/s
B - 10/s E - 100/s
C - 20/s * - < 100 b/s
1 - Each second

AD-A090 087 NAVAL OCEAN SYSTEMS CENTER SAN DIEGO CA F/G 17/2
AIRCRAFT FIBER-OPTIC INTERCONNECT SYSTEMS PROJECT.(U)
AUG 80 R D HARDER
UNCLASSIFIED NOSC/TR-576



4.3.2 PROCESSING SUBSYSTEM ARCHITECTURES

4.3.2.1 Signal Processing Subsystem

Signal processing is an array operation. General-purpose processors tend to operate on small tables and sometimes single operands. Signal processors perform the same arithmetic operations on large arrays of data.

Signal processing speed is achieved by pipelining and overlapping entire data arrays. For example, a high-speed arithmetic operation on 512 points of data can be overlapped with the staging of the next 512 points of data in a high-speed buffer store. Most signal processing operations are highly repetitive. Consequently, the arithmetic elements can be optimized to perform complex multiply and add operations at very high rates.

Since high-speed arithmetic elements are easily achievable, and vast amounts of data must be analyzed, array processors of this type are often data limited. The process of managing the vast data arrays and then feeding the arrays at high rates past the arithmetic elements becomes a problem. The design of a signal array processor is contingent upon a solid approach to the data handling problem.

Figure 4-3 illustrates the acoustic processing string in the projected 1990 Avionics Processing System. The signal processor consists of five elements:

- An input element
- A data transfer element
- An arithmetic element
- A bulk store
- A high-speed buffer store.

The input element and data transfer element are the high-speed data movers. The bulk store is a large (approximately 1 megabyte) data store with two independent high-speed buses. The input element drives one "tail" of bulk store, and the data transfer element drives the other tail. The bulk store acts as a swinging buffer between the two data-moving elements.

The high-speed store acts as a local buffer for the arithmetic element. All data arrays to be operated on in the arithmetic element must be placed in the high-speed store by the data transfer element. The high-speed store has three tails. One tail is driven by the transfer element, one tail belongs to the arithmetic element, and the remaining tail is for data to be transferred to the signal post-processor. The high-speed store acts as a triple swinging buffer between the data transfer input, the arithmetic computer, and the data output.

Table 4-1. Avionics Processor Data Rate Estimate
Sea Control/Open Ocean/1990-2000

Avionics	Input kb	Input R	Output kb	Output R	Avionics	Output kb	Output R	Input kb	Input R
Doppler	*	*	0.3	A					
Magnetic compass	*	*	0.3	C	Bulk store	10,000	1	10,000	1
Laser	*	*	0.3	C	Buffer store	320,000	1	320,000	1
IR	7.7	D-E	2.6	D	Memory	320,000	1	320,000	1
MAD	2	D	2	D					
Acoustic	*	*	10,000	1	SO	3	C	0.4	A
Radar (ASW)	0.8	B	3,000	1	ATO I	3	C	0.4	A
IFF Interrogator	*	*	-	-	ATO II	3	C	0.4	A
					Pilot	3	C	0.4	A
Actuators	6	B-E	6	B-E	Display controllers	*	*	80	C
Engine sensors	*	*	*	*					
Controls	*	*	2.5	D-E	VHF	16	C	16	C
Radar altimeter (3)	*	*	2	C	Security	48	C	48	C
Rate gyros (3)	*	*	20	D-E	ICS	32	C	48	C
Accelerometers (3)	*	*	20	D-E	ILS	0.3	B	*	*
Air data sensors (3)	*	*	4	C	AWCLS	0.8	B	*	*
Flight control computers (3)	12	D	120	D	UHF	48	C	32	C
					IFF transponder	*	*	*	*
AAES	0.5	1	0.5	1	JTIDS	16	1	16	1
Stores management	0.7	1	0.7	1	TACAN	1	C	*	*
					Radar (AEW)	650,000	1	15	E
					ESM	36,000	1	3	C

Refresh rate

A - 5/s D - 50/s
B - 10/s E - 100/s
C - 20/s * - < 100 b/s
1 - Each second

At any one time, the signal processor flow just described can be overlapping operations on four data arrays; for example,

- Input element storing array 1
- Data transfer element fetching array 2 and storing it into high-speed store
- Arithmetic element operating on array 3
- High-speed store transferring array 4 to signal postprocessor.

On the end of an array operation, each element of the signal processor transmits a sync message to the other elements in the chain to indicate that a buffer switch is imminent.

Each array operation in the signal processing chain is described by a command word. The command words are chained together and stored in each system element's program storage. A chain of these command words forms a subroutine. These subroutines are linked together by the executive controller.

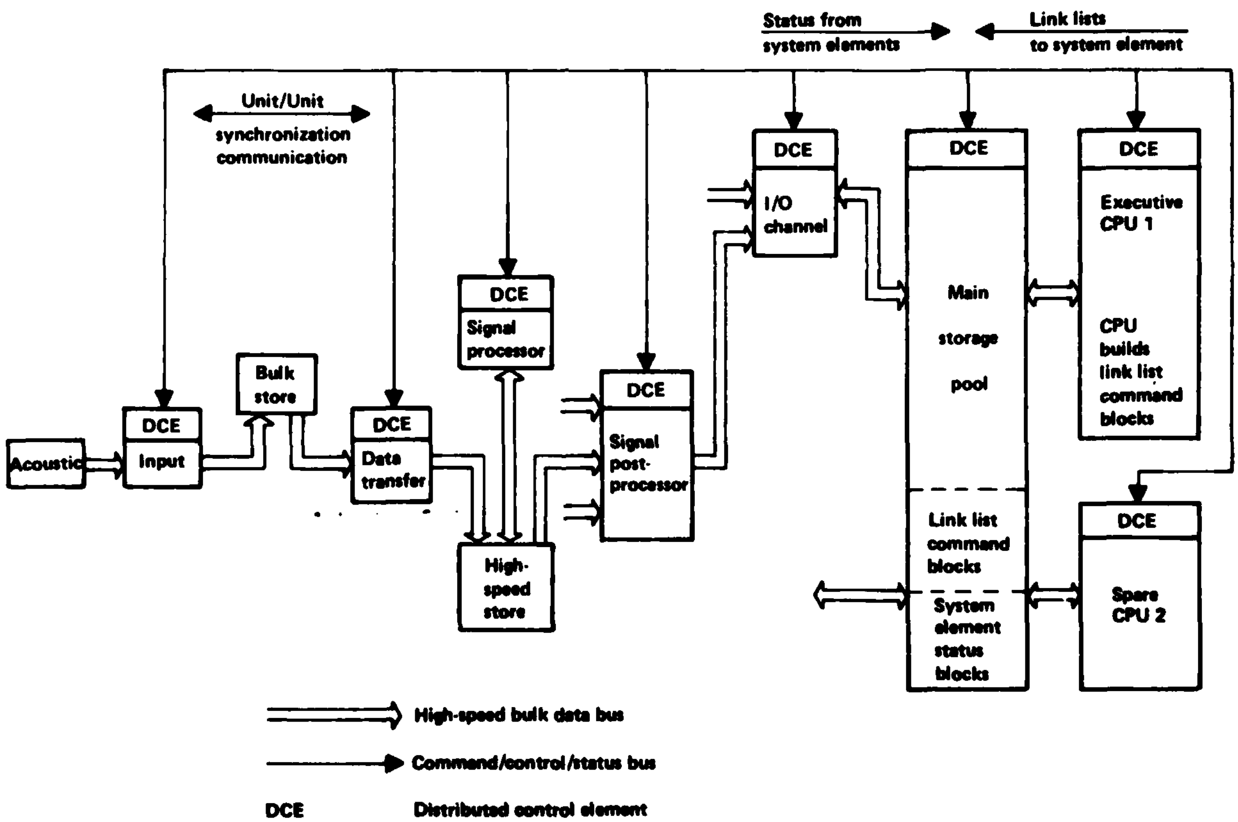


Figure 4-3. ASW Avionic System Acoustic Processing String

Once the executive controller links together array subroutines in the various elements, the system runs without further interference from the executive until a system mode change is required. The synchronization messages are imbedded in the individual element command words, and each element performs its own synchronization.

The distributed control elements (DCEs) contain the local program storage for the command chained subroutines and interface with the common distributed command and control bus. Subroutine linkages and sync messages are communicated via the distributed command and control bus. Each system element contains a DCE. Thus, all system elements have a common control interface via a standard piece of hardware.

The combination of input, data transfer, and arithmetic elements, coupled with the common DCEs, permits a common modular approach to signal processing. In the projected avionics system (see Figure 4-2), the radar signal processor and ESM receiver processor are both configured out of the same modular elements as the acoustic signal processor. Therefore, all three signal processing functions are tied into the overall system by using a common command and control bus. All signal processing elements use the standard DCE as a system control adapter.

4.3.2.2 General-Purpose Data Processing Subsystem

The GP data processing portion of the projected avionic system is illustrated in Figure 4-4.

General-purpose processors, in general, do not operate on large arrays of data. To be sure, some GP operations involve large data table operations; however, the design of these machines assumes that each operand is unique. The GP performance is usually obtained by overlapping individual instructions and operands. The GP instructions usually specify a single operation in one instruction, whereas signal processors specify an operation on an array of data in one instruction.

A GP machine in the 5-million-instruction-per-second range is projected for the future avionic system. This implies a single operation to storage can be performed in an average execution time of about 200 ns. Most processors of this type use a local high-speed history buffer known as a cache. Assuming that approximately 80% of all operands and instructions come from cache, a storage bus data rate of at least 64 bits per 400 ns is required to maintain performance. (Each instruction requires a 32-bit instruction and a 32-bit operand). This represents a data bandwidth from CPU storage of 160 Mb/s.

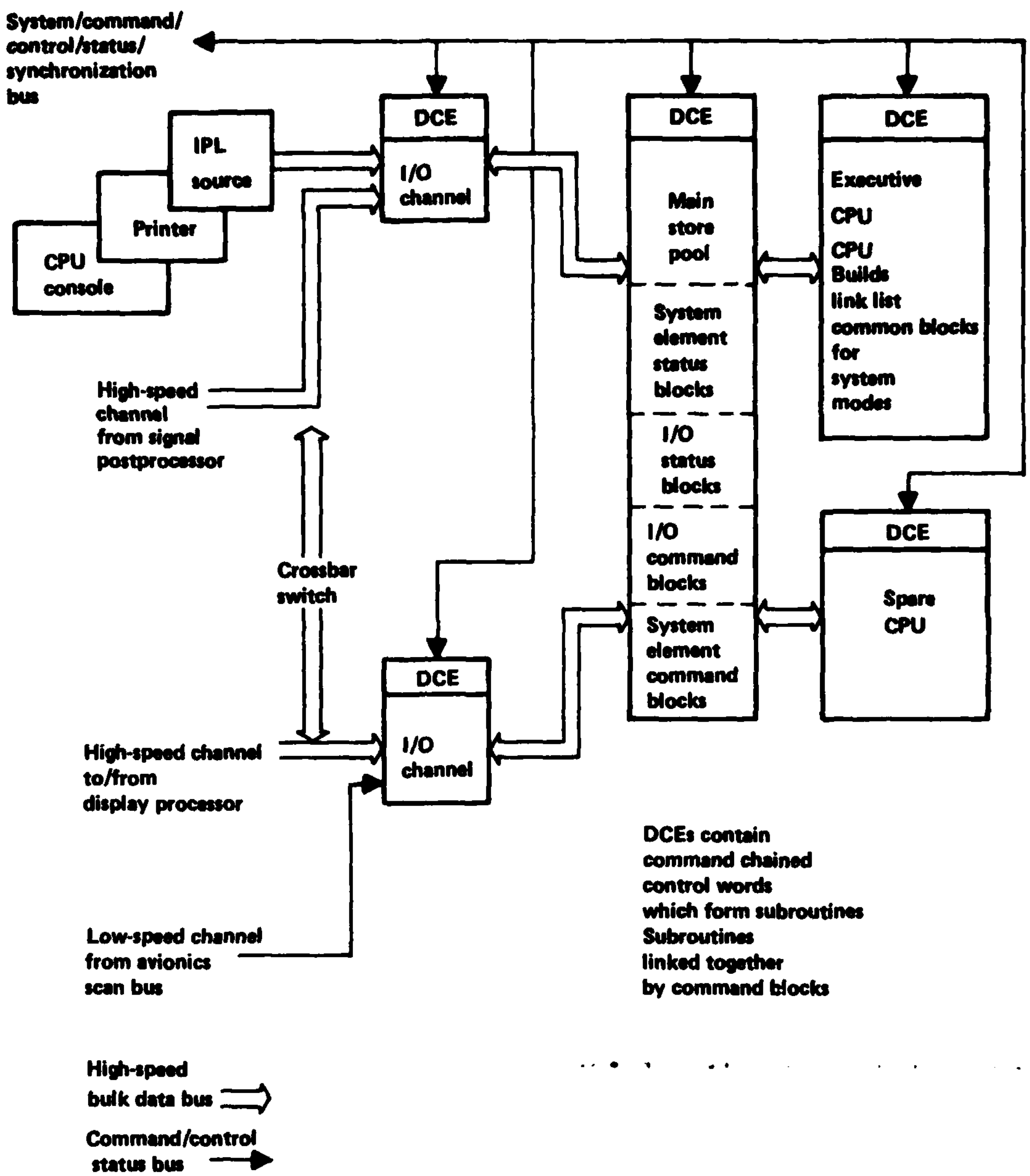


Figure 4-4. ASW Avionic System Data Processor

4.3.3 DATA HANDLING SYSTEM INTERCONNECT REQUIREMENTS

The avionic system presented in the preceding sections represents only one approach to a 1990's ASW system. Each system design, however, must consider both the control flow and the data flow in developing its interconnect requirements. Although the system design presented takes care to separate the command bus from the system data buses, these two types of data could be combined on the same bus, should the data rates of the bus not be exceeded.

This section of the report attempts to analyze the control flow and data flow requirements of the projected system and establish an upper bound on these requirements. Given an upper bound on the data flow and control flow system requirements, a data bus can then be selected for the particular system design.

Since the high-performance CPU requires significant storage data rate, there is little room for I/O interference on the storage interface. I/O interference is minimized by allowing each I/O channel its own dedicated interface to storage. Storage is further partitioned into modular independent blocks. Each modular independent block of storage has its own control logic as if it were a separate storage. It is, therefore, possible to have an I/O channel running at full rate out of one storage module while the CPU runs at full rate out of another storage module without interference. Typical storage module size is 32K x 64 bits.

All of the storage modules taken together form the main storage pool. For mission survivability, spare storage modules can be switched in automatically to replace failed modules. In case of such a failure, the system executive status is periodically checkpointed on a recording medium.

The main storage pool in the projected system has four tails: one for the main CPU, one for the spare CPU, and one tail for each of the I/O channels.

The I/O channels are controlled in the same manner as the signal processor elements. The I/O channel DCEs contain chained command words that are linked together by the executive. I/O channel command linkage, status, and synchronization messages are communicated via the DCEs onto the command/control bus.

The CPU functions to assemble link lists. These lists link together the various subroutines in the system processing elements. The CPU gathers intelligence from the system elements, makes executive decisions based on these data, and then causes system reconfigurations to occur by building the next system mode out of linked subroutines.

4.3.3.1 System Command And Control Requirements

The command and control messages communicated in the projected system fall into three categories:

- Link list command blocks
- System element status blocks
- System synchronization messages.

To establish a total system load, the load for each system element was computed independently. All of the system element loads were added to determine the total load on the command and control bus.

For example, the system load contribution from the acoustic processor arithmetic element was computed as follows:

- The arithmetic element performs eight digital filters, one FFT, and one short-term averaging process segment on 64 channels every 78 ms (13 times/s).
- Command block size = 10 process segments x 64 channels x 4 words/CMD block = 2,560 words
- Total system load = CMD block size x rate + status block size x rate + sync words x rate
- Assume the status block size equals the command block size; then, total system load = 2,560 words/block x 13 blocks/s + 2,560 words/block x 13 blocks/s + 7,168 sync/block x 13 block/s = 157.3K word/s. Because each word is 16 bits, total data rate is approximately 2.5 Mb/s.

Therefore, the acoustic arithmetic element contributes 2.5 Mb to the command and the control bus.

In a like manner, the command and status blocks, and their associated paging rates, were estimated for all of the system elements. The results of this tabulation are summarized in Table 4-2. The total peak data load for all system elements is 11.75 Mb/s.

Table 4-2. Command/Control Bus Data Requirments

System Element	Command Block Size	Command Block Rate	Data Rate 1,2 (Mb/s)
Acoustic signal processor	2,560	13	2.500
Acoustic data transfer	640	13	0.130
Acoustic input	640	13	0.130
ESM receiver signal processor	2,560	13	2.500
ESM receiver data transfer	2,560	13	2.500
ESM receiver input	640	13	0.130
Radar signal processor	64	60	0.256
Radar data transfer	64	60	0.256
Radar input	16	60	0.040
Signal postprocessor	1,200	13-60	0.600
I/O chain 1	12	13-60	0.014
I/O chain 2	80	4-60	0.164
Display processor	64	60	0.160
		Total system data rate	11.75 Mb/s

Notes: 1. Data rate computed as CMD block x rate + status block x rate + sync words x rate

2. Status blocks and sync words not shown on this chart

Once the command block sizes and the associated paging rates have been established for the system elements, it is possible to estimate the average command block size for the entire system. In this case, the average command block size is 500 words. The tabulation of the average block size is shown in Table 4-3.

What emerges from this analysis is a data bus which is capable of approximately 12 Mb/s information rate in which 80% of the messages are blocks of 500 words, and the remaining 20% of the messages are short, one-word sync messages.

Later sections of this report will show that the proposed 50-MHz optical data bus will meet the information requirements described above.

Table 4-3. Command/Control Bus Message Length Requirements

System Element	Command Block Size	X	Blocks in 1 second		
Acoustic signal processor	2.50 (K words)		13	=	32.5K
Acoustic data transfer	2.50		13		32.5
Acoustic input	0.60		13		7.8
ESM signal processor	2.50		13		32.5
ESM data transfer	2.50		13		32.5
ESM input	0.60		13		7.8
Radar signal processor	0.06		60		3.6
Radar data transfer	0.06		60		3.6
Radar input	0.01		60		0.6
Signal post processor	1.20		60		72.0
I/O Channel 1	0.01		60		0.6
I/O Channel 2	0.08		60		4.8
Display processor	0.06		60		3.6
			498 blocks	of	234K words

Total system bus load in 1 second consists of

498 command blocks of 234K words

498 status blocks of 234K words

117K (approx) syncs of 117K words

$$\text{Average block transfer} = \frac{\text{468K words}}{\text{1K messages}}$$

Average block transfer = 500 words

Average sync transfer = 1 word

$$\frac{468K}{585K} = 80\% \text{ of all messages are blocks}$$

4.3.3.2 High-Speed Bulk Data Bus Requirements

In the projected avionic system block diagram, Figure 4-2, the heavy double lines represent bulk data buses. In today's technology, these buses are highly parallel, synchronous interfaces that transmit vast amounts of data. As new interfacing technology becomes available, such as fiber optics, these buses will and should be replaced by faster, more noise tolerant devices.

System storage requirements are tending to increase well into the megaword range. With such large storage requirements, the physical distances between storage elements is becoming prohibitive. It is difficult with today's devices to achieve high speed over such large storage arrays.

In present technology, both in signal and GP processors, the heaviest data flow requirements appear to be in the storage area. To establish an upper bound on the bulk data bus requirements, two of today's processors will be analyzed for storage interface requirements, and these results will be extrapolated to form the 1990 system requirements.

Figure 4-5 illustrates the storage organization of the AN/UYS-1 (Proteus) Advanced Signal Processor. Eight 16K by 72-bit storage modules are interleaved by low-order storage address bits. This interleave causes the eight 800-ns storage modules to appear as one 100-ns storage. Of the 72 bits, only 64 are data. The remaining bits are check and parity bits for single error-correcting, double-error-detecting code. The total data rate is 720 Mb/s.

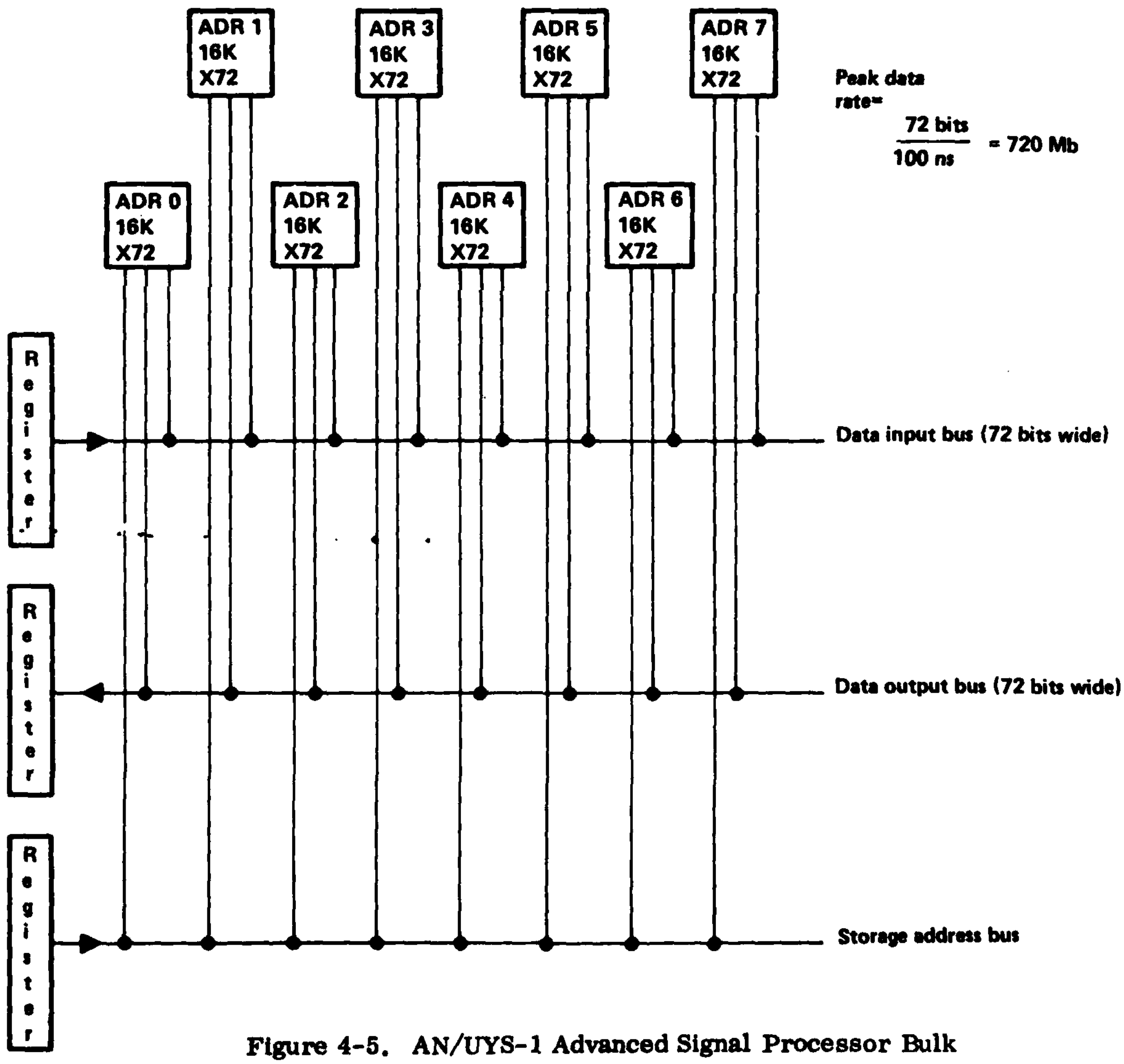


Figure 4-5. AN/UYS-1 Advanced Signal Processor Bulk Storage Organization

Figure 4-6 illustrates the storage organization of the main storage storage pool for the CC-2 GP processor used in the NATO E-3A AEW avionic system. In this system, there is no interleave. However, each core memory unit (CMU) has four independent interfaces.

The four interfaces allow the CPU and I/O to fully overlap out of different storage blocks. Only five CMUs are required to run the system; the remaining CMUs can be automatically configured into any address space, should a failure occur. Of the 72 bits, 64 are data; the remaining 8 bits are simple byte parity. The data rate for one CMU interface is 90 Mb. Because the E-3A storage pool is physically large, the interface noise problems were optimized by using a completely impedance matched 90-Ω coaxial system.

Note that the signal processor storage data rate is high to accommodate many time-multiplexed users. The signal processor data rates would be more "reasonable" if the bulk store has several independent interfaces, as does the E-3A GP machine. For example, if the signal processor storage had four tails, rather than one, the data rate per tail would be 180 Mb.

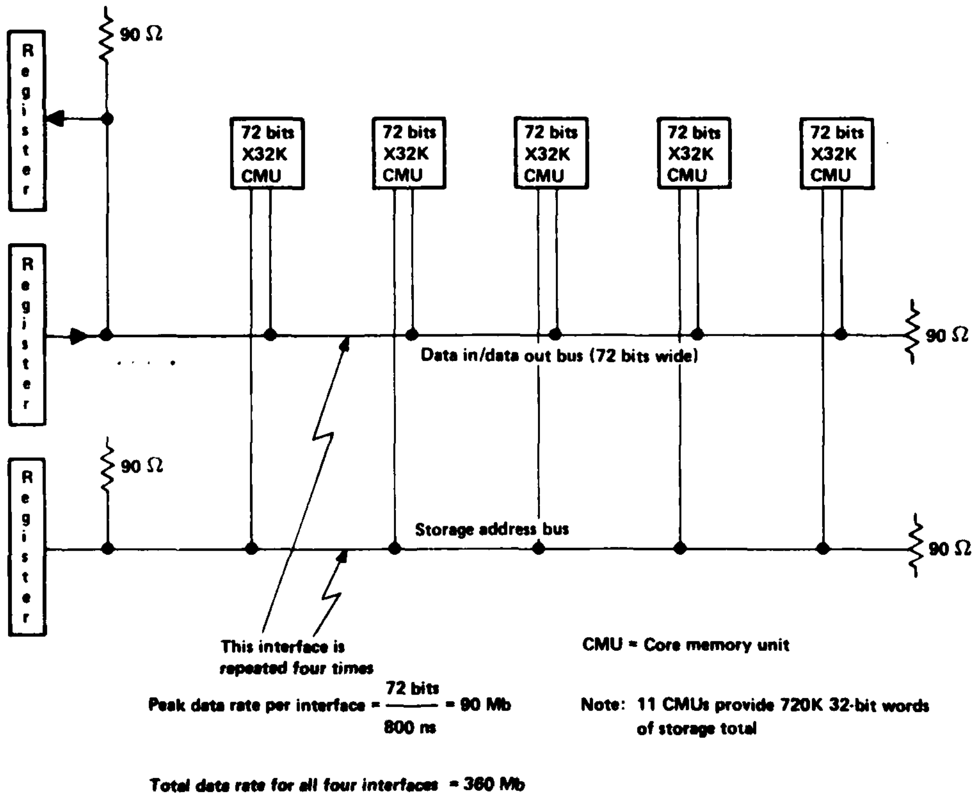


Figure 4-6. NATO E-3A Storage Pool Organization

The CC-2 GP processor is a 1.0- to 2.0-million-instruction-per-second (MIPS) machine. The AN/UYS-1 Signal Processor is a 10-million-multiply per-second (MMPS) machine. The projected GP processor should be a 5.0-MIPS machine. The projected signal processor will be in excess of 20 MMPS. By applying a simple linear extrapolation, one can estimate the projected data rates for the 1990 bulk data bus. These projections are shown in Table 4-4. The upper bound on a bulk data bus would be in the range of 1.5×10^9 bit/s.

Table 4-4. Bulk Data Bus Requirements

	Bulk Data Rates (Mb/s)	
	Current	Projected
Signal processor		
Single-tail storage	720	1440
Four-tail storage	180	360
GP processor		
Single-tail storage	360	720
Four-tail storage	90	180

4.4 CANDIDATE INTERCONNECTION SYSTEMS

In the previous section, a 1990's Avionic System was presented. This projected complex of processors was analyzed, and system data-handling requirements were developed for two categories of system communication:

- System command, status, and synchronization
- Bulk data transfers.

Both types of system communication can be combined on one bus as long as the bus data rate is not exceeded and the system requirements are satisfied. However, care should be taken in the system design to prevent the urgent system synchronization messages from being "locked out" by a long-running data-transfer block.

This section of the report functionally describes two data buses that meet the requirements developed.

4.4.1 DISTRIBUTED COMMAND AND CONTROL BUS

The distributed command and control bus is the main system control link. This bus should meet the following system requirements:

- Information rate: 12.0 Mb/s
- Average block transfer: 500 words
- 20% of all transfers are short sync messages
- Multiport, multiaccess
- Dual redundant.

This bus provides the main path for the communication of system link list command blocks, system element status blocks, and system synchronization messages.

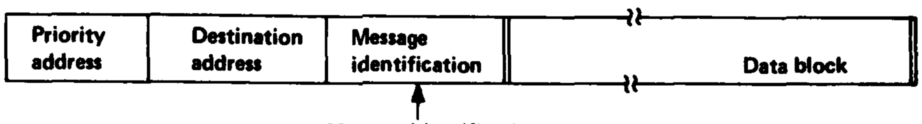
Because system synchronization messages are urgent, they should be afforded the highest priority. The system synchronization is decentralized, each system element broadcasting its "post" synchronization message to other waiting system elements. The decentralized nature of the system synchronization requires a multiport, multiaccess, distributed bus.

4.4.1.1 Distributed Bus Organization

In Figure 4-2, each system element has a port on the command and control bus. There are 16 ports total. Each port can access the bus through its DCE. Each port has a priority number consisting of 4 bits. The ports which must transmit short synchronization messages are assigned the highest priority numbers. (Priority resolution is discussed in paragraph 4.4.1.2).

Figure 4-7 illustrates the message format, which contains three subfields:

- Priority subfield is a 4-bit binary address used in resolving bus request priority. Hexidecimal F(1111) has highest priority and 0(0000) has lowest priority.
- Destination address is a 4-bit binary address which identifies the destination of a message. The destination address is the same as the priority address of the system element receiving the message.
- Message ID is a 4-bit field which identifies the message type. There can be 16 different message types. The different message types are shown in Figure 4-7.



Message Identification

0	Broadcast system reset	3	Control word	7	Sync message
1	System element reset	4	Interrupt word	8	Request element status
2	System element IPL	5	Load command block	9	System element
		6	Store status block	A	System element busy

Figure 4-7. Distributed Bus Message Format

All transfer sequences that occur on the bus require the exchange of the standard fields just described. This is required to resolve priority, alert the destination address, and identify the message being transferred. Some transfer sequences such as sync messages are immediate; that is, all the required information is contained in the standard fields. For long block transfers, the data is transmitted immediately following the standard fields.

For example, a typical sequence on the bus could consist of Priority 2 (the CPU) talking to Priority D (the acoustic arithmetic element) with a control block load (message ID) of 2,500 words.

It is proposed that this bus be implemented by a 50-MHz optical serial Manchester-encoded bus. Later sections will show that the priority resolution and standard subfield overhead imposed on the 50-MHz basic data rate will not limit the bus information rate below the 12-MHz system requirements.

4.4.1.2 Distributed Bus Access and Control

4.4.1.2.1 Distributed Priority Resolution

For a bus to be distributed, no central method for arbitrating contention can exist. Each port of the bus must contain the logic for examining all of the requests for the bus and then determining if the bus is available. A simple parallel solution would consist of each port receiving the requests of all of the remaining ports. If no higher priority ports are requesting, the port could then use the bus.

Because the proposed system synchronization is not centralized, the system sync messages should be communicated in a distributed manner. The proposed optical bus is a 50-MHz bit rate, serial, Manchester-encoded bus. One of the characteristics of an optical bus is that two light sources have a Boolean logical OR relationship. This characteristic can be exploited to form a priority resolution scheme. The following paragraphs discuss two bus use contention schemes.

4.4.2.2.1.1 Overlay Priority

In the overlay priority scheme, each user of the bus is assigned a 4-bit binary priority number. Priority 1111 (hexidecimal F) is highest priority and priority 0000 (hexidecimal 0) is lowest priority. This priority scheme uses the logical OR property of the light bus. That is, if two light sources are transmitting a binary bit onto the bus bus, the bus will go to "1" if either light source is on; consequently, logical "1s" override logical "0s" (see Table 4-5).

Table 4-5. Overlay Priority Scheme

Light Source 1	Light Source 0	Resulting Bus Level
0	0	0
0	1	1
1	0	1
1	1	1

To accomplish priority resolution, each user of the bus gates the most significant bit of the priority number onto the bus and then examines the bus. If a user gates a "0" onto the bus and the bus went to a "1", that user is eliminated from the use of the bus and does not participate in the remaining three frames. Users gating logic "1s" onto the bus are allowed to proceed to the next frame. If a user gates a "0" onto the bus and the bus remains a "0", the user proceeds to the next frame. This procedure is repeated for all 4 bits of the priority number with logic "0" users being "knocked off" by logic "1" users. The user which survives all four frames of the poll takes control of the bus.

Figure 4-8 illustrates a sample poll between four contending users. In the first frame of the poll, all four users survive. In the second frame, user 8's "0" is overridden by users D, E, and F's "1", and 8 drops off. This process is repeated until user F has won the poll at the end of four frames. the jagged line indicates the users that are participating in the poll for each frame.

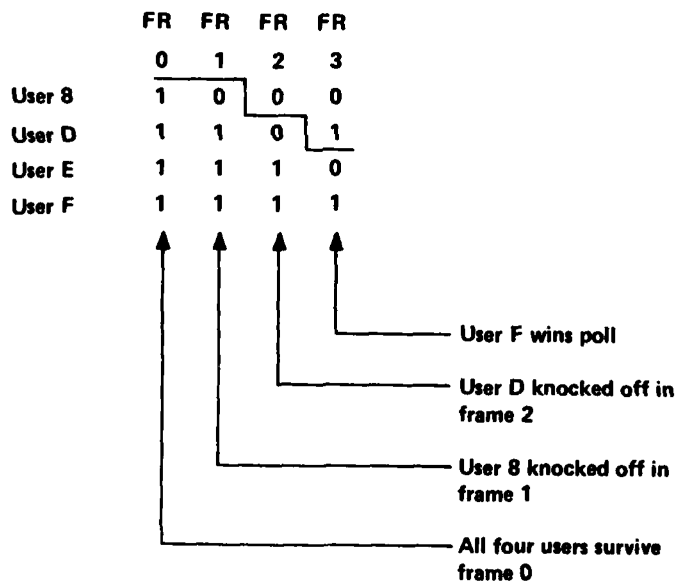


Figure 4-8. Sample Overlay Priority Poll

This type of priority resolution is quite efficient and lends itself well to serial applications. For example, 16-way priority can be resolved in just four frames. 2^N way priority can be resolved in N frames.

4.4.1.2.1.2 Timeout Priority

In the timeout priority scheme, each user of the bus monitors the bus quiet time, which is the gap between successive clocks on the bus. A minimum quiet time is chosen to indicate that the bus in not in use. The minimum quiet time must be greater than the travel time between the two furthest ports on the bus.

The highest priority device is allowed to take control of the bus when the bus has been inactive for one quiet time, Tq. In the process of taking control of the bus, the highest priority user develops a clock which resets the quiet timers in the other devices. This alerts the other ports that the bus is not available to them.

Each device on the bus keeps a quiet timer. Priority 2 must wait two quiet times before it can take the bus: one quiet time to determine if the bus is available and another quiet time to determine if Priority 1 is going to take the bus. Therefore, the device priority is determined by the number of wait or quiet times the device must wait.

For a 16-port bus there are 16 quiet times, with an average priority resolution time of eight quiet times. For N devices, the average priority resolution is N/2 quiet times. This approach is less efficient than the overlay priority. Overlay priority requires log 2^N priority frames for N devices; or four priority frames for 16 devices.

4.4.1.2.2 Default Demand Response

Although the proposed bus is distributed, enough additional message IDs have been provided in the message format to allow the bus to default to demand/response mode of operation.
Message ID codes are as follows:
- 8 — Request Element Status
- 9 — System Element Available
- A — System Element Busy.

These codes have been included in the format (see Figure 4-7) to allow a central controller to interrogate a remote device. For example, a central controller can force a remote device to give status by transmitting
- Request Element Status.

The system element can then respond with:
- Store Status Block

  or
- System Element Busy.

In addition, message ID 3, Command Word, is designed to transmit one 32-bit word which can be further interpreted by the particular device. The device can respond System Element Available Or Busy, depending on the device status to the command word.

By including demand/response as a subset of the overall bus, one bus can be developed for *multiple system applications.*

4.4.2 HIGH-SPEED BULK DATA BUS

High-speed bulk data applications, such as CPU storage pools or signal processor bulk storage, are projected to require data rates in the range of 180 to 1,440 Mb/s.

Current TTL Schottky technology operates at a clock rate of about 10 MHz. This clock rate is limited by parasitic electrical effects such as cable capacitance and ground return path inductance. These parasitic effects cause noise problems which manifest themselves as long data settling times on the data buses. At the 10-MHz clock rate, the vast data rate is achieved by a lot of parallelism. As many as 72 lines are used in parallel to achieve the data rates.

Optical data buses are not affected by the parasitic noise effects of capacitance, ground return inductance, and crosstalk. They provide a noise-tolerant, high-speed method for bulk data transfer.

This report recommends the development of a 50-MHz distributed optical data bus. By paralleling the optical lines, one can gain a 5:1 data rate advantage over today's 10-MHz buses.

The reliability of the parallel optical bus can be improved by using additional parallel lines for error correcting code. With error correcting code, a failure in one line is corrected by the check bits carried on the additional lines. If the probability of a failure on one line is A, then the probability of a failure on eight lines with error correcting code is A^2, because two lines must fail to take out the system.

Table 4-6 illustrates the number of parallel optical data lines required to meet the data requirements outlined in paragraph 4.3.3.2.

Table 4-6. Parallel Optical Data Lines

	Required Data Rate (Mb/s)	Optical Bus Data Lines	Optical Bus Check Lines
Signal processor			
Single-tail storage	1440	32	6
Four-tail storage	360	8	4
GP processor			
Single-tail storage	720	16	5
Four-tail storage	180	4	3

An eight-line optical data bus with four check lines would provide a 400-Mb/s data bus capability. This bus would cover all the multiple tail storage applications.

It is recommended that a parallel synchronous optical data bus be developed for bulk data applications.

4.5 PROPOSED INTERCONNECT SYSTEM IMPLEMENTATION

Based on the trades and analyses of previous paragraphs, Table 4-7 gives the performance goals for an optical data bus.

Table 4-7. Optical Data Lines, Performance Goals

Bit rate (serial):	50 MHz
Number of terminals:	16 minimum
Maximum terminal separation:	100 m
Bit error rate (undetected):	1×10^{-12}
Timing:	Asynchronous
Protocol:	Free access
Priority resolution:	Timeout priority

These goals were chosen as reasonable and achievable during the task IIa contract phase.

The 50-MHz bit rate was chosen because of the previously discussed protocol inefficiencies and gives sufficient margin for system growth. The 16 ports on the bus and 100-m terminal separation seem adequate for the majority of avionic systems (large and small). A bit error rate of 1×10^{-12} should keep the work or message error rate well within system requirements for reliable operation.

The Timeout Priority scheme was chosen for simplest electronics and would be adequate for the demonstration hardware. Although the Overlay Priority scheme was more efficient, it required simultaneous operation of many transmitters and could lead to receiver overloads and a requirement for much greater receiver dynamic range.

4.5.1 OPTICAL MODULATION AND WAVEFORMS

Without some method of master clock distribution, data on the bus must contain its own timing or clock information. For this reason, biphase Manchester-encoded data will be assumed at the electrical/optical interfaces at the transmitter and receiver ends of the terminals. Selection of an optical modulation method depends upon the capability of available technology and an achievable signal-to-noise ratio for the required bit error rate (BER). The method requiring the least bandwidth in both driver and receiver is the direct modulation of an emitter with the Manchester-encoded data. Although better signal-to-noise ratios may be achievable with other modulations, the bandwidth requirements would complicate the design and result in much more complex and less reliable circuitry. Direct on-off modulation with the Manchester data was therefore chosen.

4.5.2 SOURCES AND DETECTORS

4.5.2.1 Sources

Sources for a 50-MHz Manchester-encoded optical signal will be required to have an output bandwidth of at least 100 MHz. Semiconductor injection lasers fit this requirement easily, but require more complex temperature compensation to regulate the optical power out over the temperature range. Today's better LEDs have risetimes in the 2- to 3-ns region and can therefore also fit the requirement if care is taken in design of the driver and the receiver circuitry. The RCA 30119 or the RCA C30133 LEDs are examples of available devices. It can be safely assumed that equal or better LEDs will be available in the 1985-1990 time period.

Measurements made on purchased devices during Task IIa did not validate vendor data sheets, and an injection laser was required as a source.

4.5.2.2 Detectors

Although avalanche photodiodes (APDs) easily meet the bandwidth requirements for the detector, so do available PIN photodiodes such as the HP 4207. Selection of a PIN diode will negate the complex temperature compensation and high voltage required for the APDs but may lose 5 to 10 dB of gain in trade. With careful optical design, the PIN can achieve adequate margins.

4.5.3 ENCODING AND DECODING

4.5.3.1 Encoding

It is assumed that the digital logic circuitry in a terminal will encode the binary data to be transmitted into biphase Manchester before interfacing with the optical transmitter circuitry. The latter will directly modulate the LED.

4.5.3.2 Decoding

The optical receiver (described in Paragraph 3.6) will output logic-level Manchester data to be decoded by the logic circuitry in the terminal. Decoding of this data into "1s" and "0s" will be accomplished by utilizing a 100-MHz clock in the receiving terminal logic circuitry. Data will be fed into four-word-long shift registers, each of the four being slightly delayed (1/4 bit time) from each other. Each will be strobed or clocked by the 100-MHz local clock and tested for valid bits, sync, and parity. Whichever register "tests good" will be used for shifting the data out in parallel to be used by the terminal.

4.5.4 OPTICAL DRIVER

For an average 50% duty cycle waveform such as the biphase Manchester, peak currents from 50 to 200 mA will be required to drive the LED. To achieve risetimes on the order of 2 to 3 ns, a form of current switch will be utilized as the driver. Figure 4-9 schematically depicts the circuit.

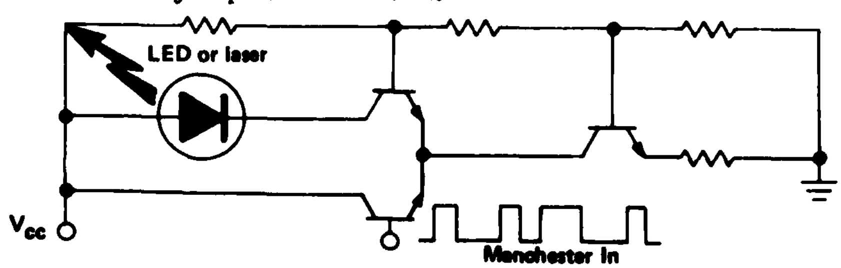


Figure 4-9. Optical Driver Schematic Diagram

4.5.5 OPTICAL RECEIVER

Two problems confront the receiver designer for a multiterminal bus system using Manchester-encoded data. An AC-coupled amplifier normally causes the average level of the output to settle to different levels, depending on the length of the message and the time between messages. This requires that the threshold level of the 1/0 detection circuit change at the same rate to keep the signal peak-to-threshold ratio a constant. AC coupling is required to circumvent the problems caused by drift in a DC-coupled amplifier.

The large dynamic range in a system where two conversing terminals may be a minimum of 1 m to a maximum of 100 m apart and may have a varying number of connector losses will aggravate the preceding problem and may cause a problem of its own in the receiver. If the gain is sufficient for the terminals far apart, circuitry may become saturated, causing possible waveform distortions, when the terminals are close together. Figure 4-10 is a block diagram of a proposed solution to both problems.

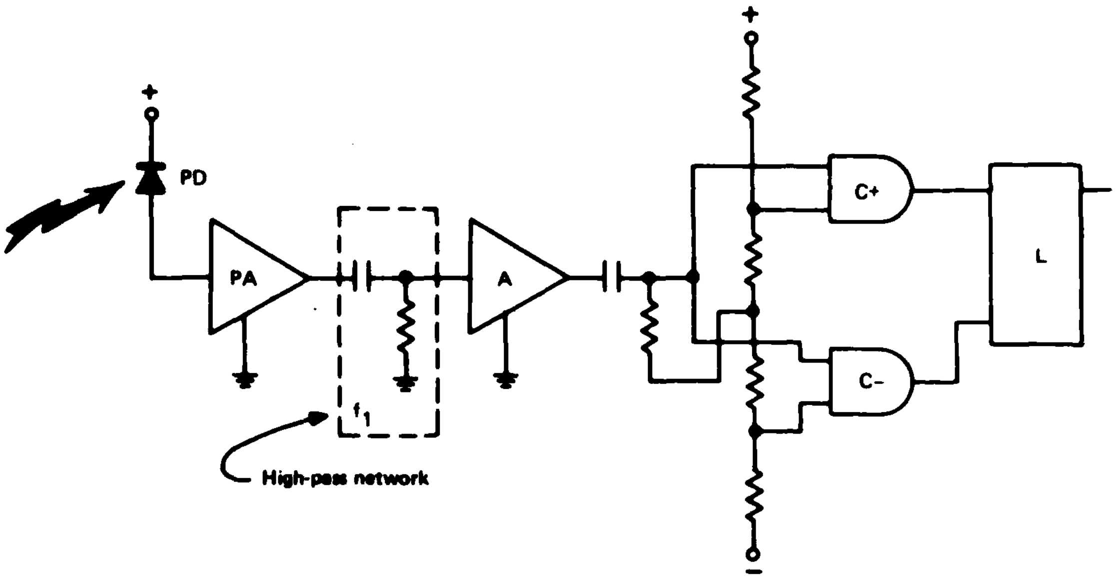


Figure 4-10. Optical Receiver Block Diagram

To maintain as much control as possible over the jitter and transition time ambiguity of the reconstructed Manchester electrical signal at the output of the receiver, it was decided to use the edges of the optical waveform (rise and fall transitions) rather than simply amplifying it linearly. The reason for this is apparent from Figure 4-11. Because of the large dynamic range required of the receiver, there must be some form of limiting for large signals. Automatic gain control (AGC) at these speeds is not feasible because of the response times needed to control the amplitude of the first pulse in a message.

In Figure 4-11A, a pulse waveform with 50% duty cycle, if amplified linearly, will have equal widths, x, and y, at the half-amplitude point. If the next amplifier limits at level L in the top waveform, the output of this amplifier will look like Figure 4-11B. Obviously, the waveform symmetry has been lost, making decoding less accurate.

To avoid this effect, a receiver amplifier concept was used wherein the low-frequency response was highly limited, and the waveform was essentially "differentiated". See Figure 4-12, where the dotted curve is the result of passing the solid curve through a high-pass filter with a pole at 35 MHz. In this case, further amplification does not as severely affect the location in time of the leading edges of the waveform, even if there is limiting.

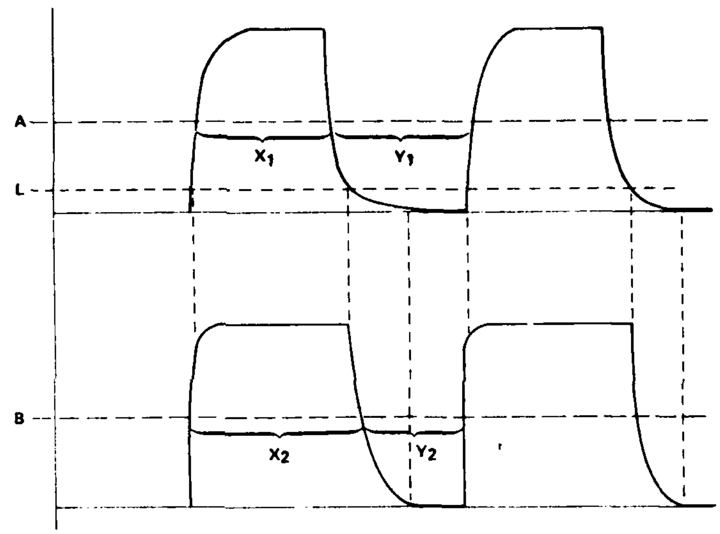


Figure 4-11. Pulsewidth Distortion Caused by Amplitude Limiting

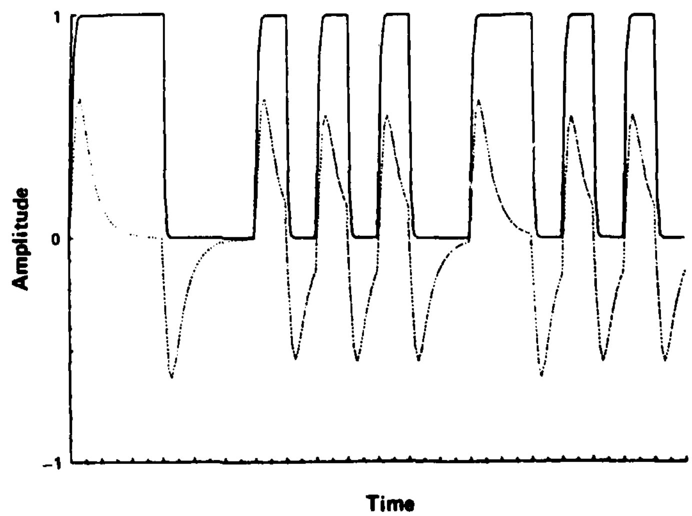


Figure 4-12. Result of Signal Passing Through a High-Pass Filter

Figure 4-13 shows the method used to reconstruct the original waveform. The threshold levels (+ and -) are for comparators whose outputs set (+) and reset (-) a high-speed latch when the input waveform crosses the thresholds. In this way, the jitter and edge ambiguity of the waveform is minimized. Figure 4-10 is the schematic diagram of this receiver.

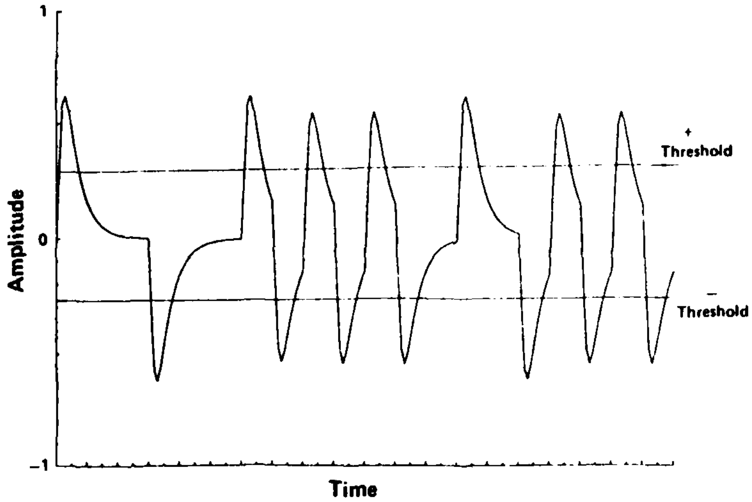


Figure 4-13. 35- to 200-MHz Signals

4.5.6 OPTICAL CABLE AND CONNECTORS

The high-speed LED sources have small junction or emitting areas and are therefore very compatible with single-fiber transmission cables. Optical fibers are generally specified for bandwidth by a frequency-length parameter (usually megahertz-kilometers). The limitation is caused by material and/or modal dispersion of the fiber/source combination. To minimize the input coupling loss and connector loss, the largest practical fiber diameter should be chosen. For 800- to 900-nm wavelength sources, step index plastic-clad fibers with 200-μm cores can be procurred, with a 3-dB optical bandwidth of 25 MHz-km and an attenuation of 5 dB/km (Quartz Products Corp. QSF-A-200). If our bus maximum run is 0.1 km, the bandwidth of the transmission medium will be 250 MHz.

Connectors compatible with this fiber can be procured today with average losses of 1 to 1.5 dB.

Work is proceeding at various vendors and agencies on sources and detectors in the 1,300-nm region of the optical spectrum. At this frequency, dispersion in glass fibers tends to be minimum. Future availability of such sources and detectors will increase the bandwidth margins and decrease the actual attenuation losses of the system.

It is therefore proposed that the transmission medium be a 200- μm core, step index, plastic-clad fiber cable, plastic cladding being utilized for maximum flexibility and minimum bending radius. As will be seen in Section 5.3, problems with termination procedures for plastic-clad fibers forced use of a glass-clad fiber for the demonstration of Task IIa.

4.5.7 OPTICAL BUS COUPLERS

For a 16-port optical data bus, it has been well established that a *radial coupler is preferred* over a series of "T" couplers to minimize the coupling loss. Most, if not all of the single-fiber couplers developed to date have been packaged with the connection means to each transmission cable as part of the coupler structure itself.

It is recommended that a coupler be developed with arbitrarily long unterminated transmission cables (say 10 m) as part of the coupler structure itself. This will allow custom installation of the coupler and avoid up to two connectors per path. It is also proposed that the coupler be transmissive rather than reflective to avoid the extra 3 dB loss at the terminal caused by the requirement to divide the path at the terminal between the source and the detector. Figure 4-14 is a sketch depicting such a coupler. It will allow the minimum package size, because only the fiber cable ends and the integration or mixing rod are in the structure.

4.5.8 BUS PROTOCOL

A free-access bus requires a protocol to determine who gets the bus during conflicts caused by simultaneous requests. This in turn requires that the terminals using this protocol be "smart" to avoid a "central" controller who determines this priority and assigns the bus. There are several methods for each terminal to determine when it can get access to the bus. The more efficient this protocol becomes, the more complex the terminal logic.

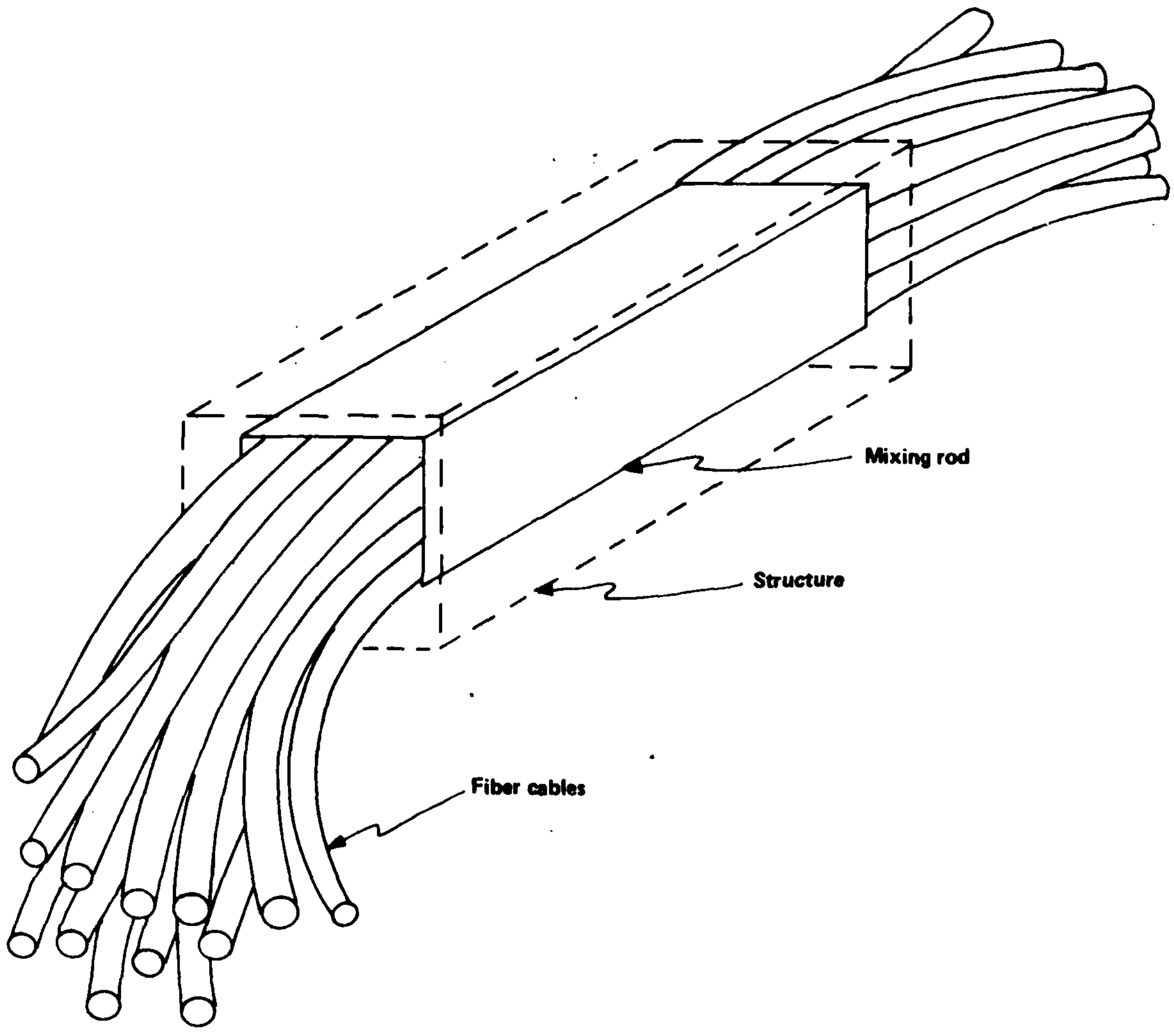


Figure 4-14. Sixteen-Port Transmissive Coupler

A further complication arises as the bus bit rate increases and/or the distance between terminals gets greater. The reason for this is that the optical signal transit time between the most distant terminals becomes a larger part of the total transaction time relative to the actual message, especially if the average message length is small. For example, at a 50-MHz bit rate, a 20-bit word takes 400 ns to be serialized and put onto the bus. If the most distant terminals are 100 m apart, it takes 1 μs for a round trip. For a 16-port serial bus, one of the more efficient priority-resolving protocols would take the equivalent of somewhere between five and six round trips to the most distant terminal and would, therefore, take 5 to 6 μs to resolve priority. A 10-word message would require about 5 μs, and bus efficiency would be around 50%, assuming all the bits in the 10 words were information.

This is the penalty paid for a free-access, high-speed serial data bus. Efficiency can be improved by buffering messages for longer average message lengths or by restricting the distance between terminals to something reasonable for the application. The most efficient bus is one with the command/response protocol. It is the one with the least logical complexity, but has also the least flexibility and does not fit in with the concept of distributed processing and control.

The protocol selected for demonstrating priority resolution for a 16-port bus requires monitoring of the bus by each of the terminals to determine when it is free. This can be done by starting a timeout in the receiver for each appearance of any pulse on the bus. Whenever the terminal senses a gap equal to or greater than the longest round trip time between the most distant terminals, it concludes that the bus is free. This gives time for a response from one terminal to the other if they had been conversing and had been the most widely separated terminals. Each terminal has been assigned a priority from 0 to 15, with 0 being the highest.

If the one-way transit time between the most distant terminals is t_d, then each terminal must wait t_d times its priority number before it tries to get the bus. This is because the terminal with priority (N) must wait to see if priority (N-1) wanted the bus, and terminal (N-1) might have been the most distant terminal. In summary, priority (N) terminal waits ($N \times t_d$) before it attempts to get the bus.

To establish the real information rate on the bus, the following equation can be used:

$$I_r = \frac{16M\ f_b}{22M + P + \beta\,(N-1)t_d f_b + 2t_d f_b}$$

where M is the average message length in words

P is the number of protocol bits

N is the number of terminals on the bus

f_b is the bus bit rate

t_d is the one-way delay between most widely separated terminals

β is a factor (<1) which determines average wait time for the average terminal

($\beta N t_d f_b$) is the average equivalent number of bits for establishing priority

$2t_d f_b$ is the initial wait time to determine when the bus is free

For $f_b = 50$ MHz

$P = 3$ words (66 bits)

$N = 16$

$$t_d = \frac{Ln}{3 \times 10^8}$$

where L is longest length in meters, n is index of refraction of fiber (1.5), 3×10^8 is speed of light in vacuum in meters per second

$\beta = 0.5$ (average wait is 7.5 priority times)

$$I_r = \frac{36.36}{1 + \frac{3 + 0.072L}{M}}$$

Figure 4-15 is a graph showing the relationship between real information rate and message length for various bus lengths. This graph does not consider conflicts where a terminal with higher priority takes the bus from one of lower priority. It is, therefore, optimistic in terms of real information rates.

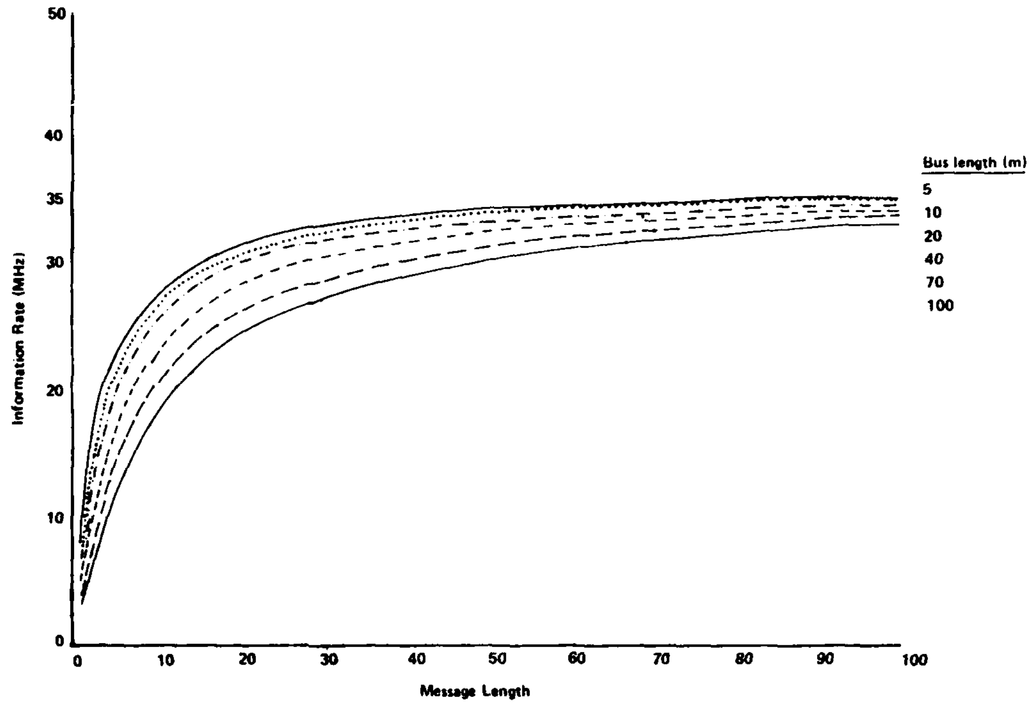


Figure 4-15. Real Information Rate For 50-MHz Bus

The information rate requirement set forth in Section 4.3.3.1 was 11.75 Mb/s, and the average message length was 5.3 words. The graph of Figure 4-15 shows that this is just within the capability of a 100-m bus, assuming no conflicts. To determine the conflicts, the actual system must be simulated and run while actually collecting data on the number and frequency of conflicts. This is so system dependent that many different possible systems must be simulated to obtain a real set of reliable data.

During Task IIa performance, some simulations will be run and data collected. In the interim, the margins can be achieved, if necessary, by restricting the bus length.

Other priority resolution schemes may require as much time, but eliminate the possibility of conflict. Such a scheme is superficially described in Section 4.4.1.2.1. What is important here, is that a 50-MHz serial data bus can satisfy the great majority of future requirements for the 1990 system.

4.6 TEST PLAN

4.6.1 INTRODUCTION

This section describes the test plan to which the future aircraft fiber-optic interconnect system was demonstrated and evaluated. This test plan may differ from the Fiber-Optic Final Demonstration System Development Plan, 79-M20-004, Data Item 001 delivered under Contract No. N00123-77-C-0747. Any differences found between these two specifications are a result of finding a more efficient and accurate means of testing during the final test and integration period. The testing was done at three levels, the component, subsystem, and laboratory system level.

4.6.1.1 Component Level Testing

All of the optical components were tested before assembly to verify that they were within their design specifications. These components included in the following:

- Source
- Detector
- Cable
- Connectors
- Multiport coupler.

4.6.1.2 Subsystem Level Testing

The transmitter and receiver circuits were connected via the optical components, and an end-to-end test was performed to verify the operation of all components.

4.6.1.3 Laboratory System Level Testing

This level of testing demonstrates the feasibility of operating a high-speed, fiber-optic data bus in a distributed control protocol system. This test consists of transferring data between terminals on a common interconnection system at a 50-MHz rate.

4.6.2 TESTING REQUIREMENTS

4.6.2.1 General

This section contains the detailed requirements, both input and output, used to verify the operation of the fiber-optic system. Separate requirements are given for each of the three levels of testing performed.

4.6.2.2 Component Level Testing Requirements

Each component was tested to verify that it meets the requirements listed. The testing was done at the operation point of the fiber-optic system for each component.

4.6.2.2.1 Source Tests

Each source was tested for total output at its published CW operating current. This was done with an integrating cube in the optics laboratory. The source then was coupled to a short piece of the cable to be used, and the light output of the cable was measured. This determined the input coupling loss because the loss of the short fiber is negligible.

The cable was then attached to the receiver through a known attenuation and the output of the transimpedance preamplifier measured. Because the transimpedance and attenuation are known precisely, the loss due to the output coupling in combination with the photodiode responsivity can be calculated. With an assumed value for responsivity, the output coupling loss can be estimated. The gain of the total receiver can be measured at this point but was deferred until later.

Because there is now a calibrated system through the receiver preamplifier, and it is linear because of the signal levels involved, this system was used to measure the output of the source with a pulse repetition rate of 50 MHz and a duty cycle of 50%.

Peak light level outputs and preamp rise and fall times were measured for each source. Because the source was driven by the actual driver to be used in the system, the rise time, peak amplitude, and fall time of the current pulse were also measured and serve as part of the total transmitter test.

4.6.2.2.2 Photodiode Tests

It is difficult to test the actual responsivity of the photodiode (PD) unless the actual light energy falling on the chip is known. Without special instrumentation, this cannot be done. As stated in the source tests, what was measured was the combination of output coupling loss and responsivity because the transimpedance of the preamplifier can be accurately measured. The responsivity assumed was that of the manufacturer's data sheet. Each of the PD/preamplifier combinations was measured in this manner.

4.6.2.2.3 Cable Testing

Once there exists a calibrated transmitter/receiver combination, the various cables can be tested for attenuation by substitution for the piece used to calibrate the pair. Measurements on a representative number were made.

4.6.2.2.4 Connector Testing

By using lengths of cable tested previously, we can measure the loss due to connectors by connecting two or more cables between the calibrated driver and receiver pair. This was done for several connector/cable combinations.

4.6.2.2.5 Multiport Coupler Testing

Using all of the preceding components, which now have known losses, the coupler was tested by alternating the transmitter among the 16 inputs and measuring all 16 outputs each time. The intrinsic loss and the variation among the outputs was then measured and recorded.

4.6.2.3 Subsystem Level Testing Requirements

After the component level testing was done, the subsystem was assembled with the optical components and the driver/receiver circuits.

4.6.2.3.1 Transmitter

With the calibrated receiver, each of the transmitters underwent tests to determine preamplifier rise time, fall time, and peak light outputs. Also, the difference in noise level (if any) between light on and light off was measured at the receiver. All measurements were at the output of the preamplifier to avoid the compression at the amplifier stages that follow, and to keep the system linear.

4.6.2.3.2 Receiver

With the calibrated transmitter, the light output was set for two different levels at the output of the preamplifier. These levels correspond to the total dynamic range expected on the input light and were accomplished by insertion of neutral density filters in the line between the driver and receiver. The signal level was measured at the input to the comparator for both settings. Peak

amplitudes and rise times were recorded. Corresponding pulse outputs from the comparator were also measured, along with their rise times, widths, and relative delays.

With the transmitter turned off, the noise (rms) level at the input to the comprator was measured. From the small signal gain measurements of the receiver, the noise referred to the input light power was calculated. This determined the receiver sensitivity.

4.6.2.3.3 Link Tests

The link tests consist of connecting the transmitter to the receiver through a short length of fiber-optic cable. A logic-level Manchester-encoded signal was input to the transmitter, and the output of the receiver was measured to verify the correct operation of the link. Minimum and maximum loss measurements were then obtained from the link. By building up from the minimum to maximum loss link, incremental losses were measured to validate and correlate to the earlier subsystem and component test results.

4.6.2.4 Laboratory System Level Testing

The system level tests were divided into two sections, the first being reliability testing, which measures the bit error rate and the incomplete message rate of terminal, the second being the efficiency testing, which measures the access time of the terminals and the protocol efficiency for the system.

4.6.2.4.1 Reliability Testing

The system's reliability is defined in terms of the bit error rate (BER) and the system's incomplete message rate (IMR). BER is defined as the number of incorrect bits received, divided by the number of bits sent. The IMR is defined as the number of messages received with errors, divided by the number of messages sent.

The original test plan called for BER and IMR measurements to be made via the IBM 5110. The IBM 5110 would generate messages of pseudo random data, program a message unit to send this data to another message unit, read the transmitted data, and then compare the data sent with that received. Several programs were written to perform this task. These programs were used extensively in testing the system. However, we found that the 5110 required too much time to analyze sufficient data to determine, with accuracy, a 10^{-12} BER. An alternate method for measuring BER and IMR allows for real-time tabulation of errors, eliminating the need for 5110 analysis.

An additional piece of test equipment was constructed with the capability of counting the number of words sent by a message unit, received by a message unit, and received with bit errors by a message unit. Figure 4-16 is a photograph of this device; the block diagram is given in Figure 4-17.

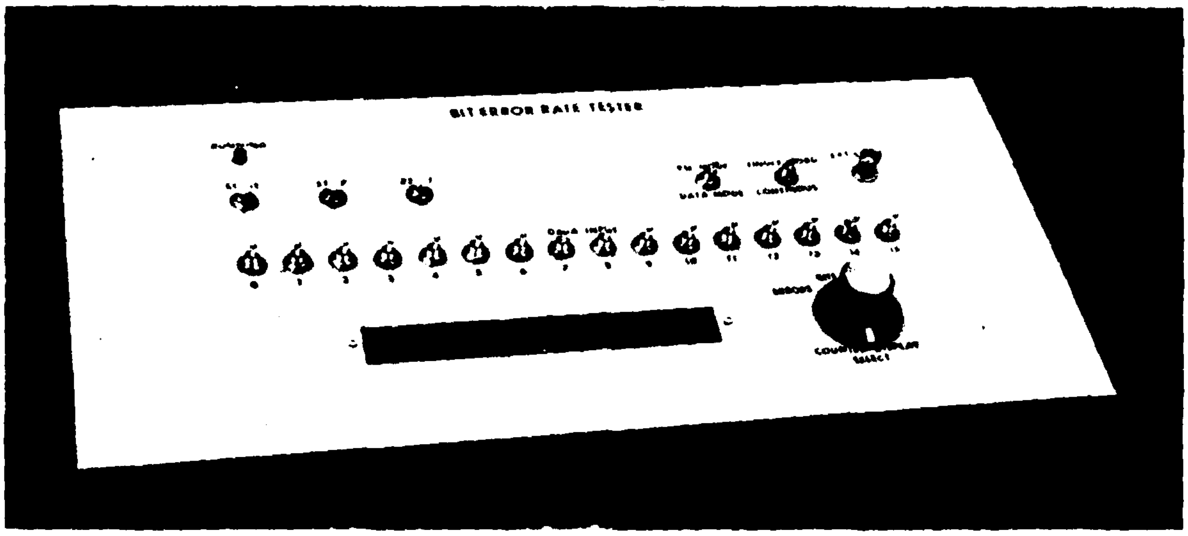


Figure 4-16. System Error Rate Counter Photograph

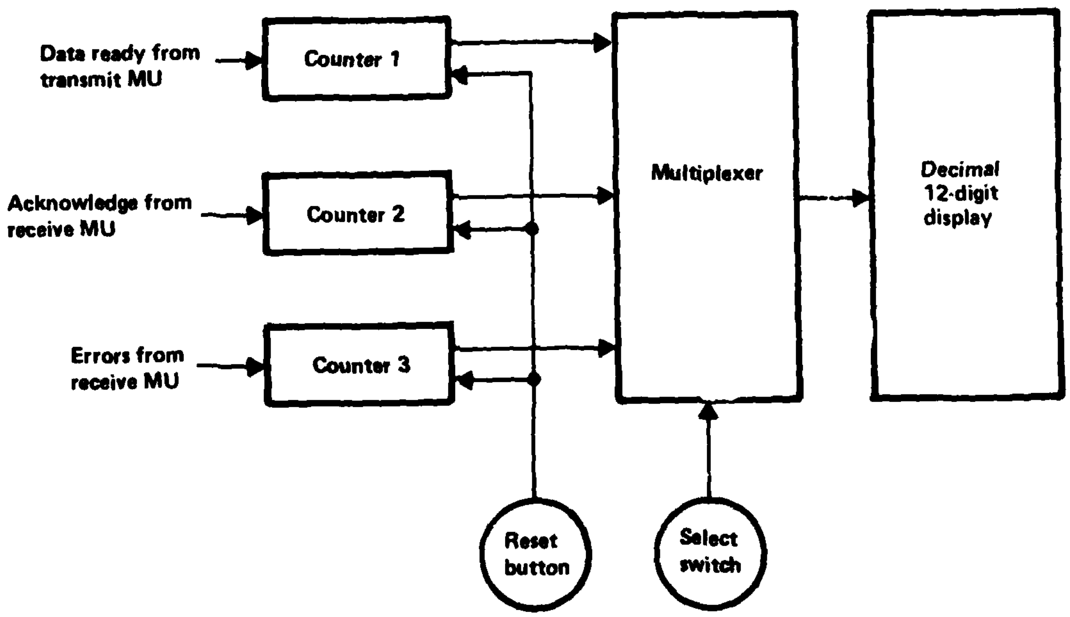


Figure 4-17. System Error Rate Counter Block Diagram

When the new piece of test equipment is attached to the system, the 5110 can program a message unit to transmit its data to a designated message unit in a repeating mode. This test was run until 1 billion words (20 billion bits) were transmitted and was performed for all nine data paths numerous times. The calculation for BER was made as previously defined, while the IMR was modified to be an incomplete word rate (IWR). The IWR is defined below:

$$\text{IWR} = \frac{\text{Number of words with incorrect bits received and number of words missed}}{\text{Number of words transmitted}}$$

The IWR is actually more meaningful because it is a function of message length, a system-dependent variable. Note that the IMR is easily determined by multiplying the IWR by the average number of words per message for a given application.

4.6.2.4.2 Efficiency Testing

The operational bus test measures the overall efficiency of transmitting information between terminals using a distributed control protocol and bus access similar to the techniques stated in Section 4.3.

The access time of a terminal is defined as the time elapsed between a terminal gaining control of the bus and the time it first requested a transmission. The access time was measured while varying quantities such as the terminal's priority number, the percentage of bus loading, and the average message length. The message units priority number, transmission rate, and average message length are parameters that can be varied by software and loaded at the start of each test.

The percentage of bus loading varies from 0% to 100% by initializing the bus loading unit with the appropriate parameters. The average message length on the bus varies from 2 words to 250 words per message. The priority number of the units is varied to determine both the true information rate for the data bus and the information rate for a terminal, with a specific priority number for different combinations of bus loading and average message lengths. Curves plotted from the tests results are message unit bus access time as a function of message unit priority, and message unit information rate as a function of message length.

4.6.3 HARDWARE REQUIREMENTS

The system level test includes the following hardware: an IBM 5110 computer and I/O control interface (message operating system), two terminals called message units, and a bus loading unit interconnected as shown in Figure 4-18. The message units and the bus loading unit are identical in design. The differences between the units are a function of the parameters loaded by the 5110 computer. The parameters loaded are as follows:

- Mode of operation
- Distributed control priority
- Message rate
- Data
- Message length.

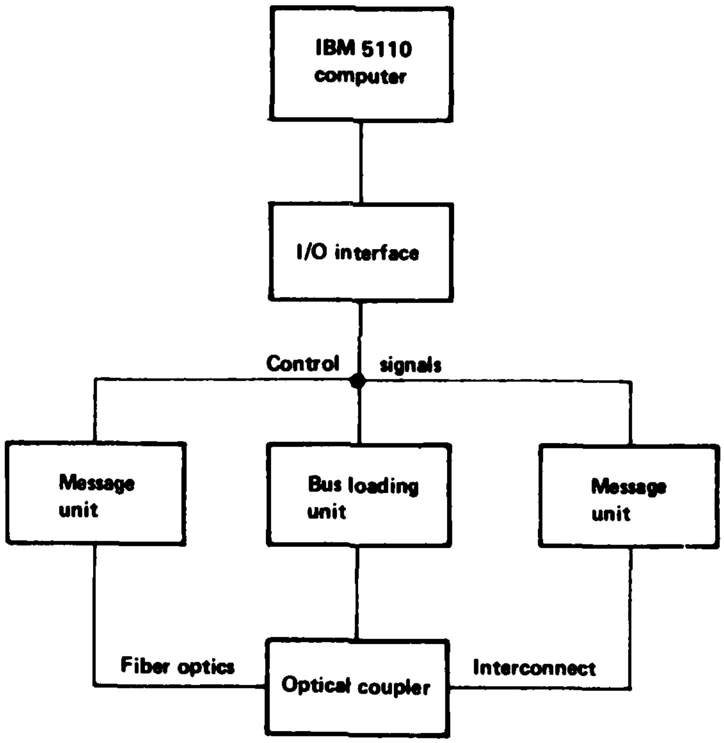


Figure 4-18. High-Speed Data Bus Block Diagram

4.6.3.1 Message Operating System

The message operating system (MOS) consists of an IBM 5110 desk top computer and an RS-232 I/O interface.

The IBM 5110 desk top computer contains software packages loaded from cartridges for message unit control, bus data generation, and bus data analysis. The I/O interface consists of a wirewrap board containing a Z80 microprocessor, memories, and control circuits that communicate with the 5110 via an RS-232C interface. The I/O interface is used to multiplex the 5110 with one of the two message units or the bus loading unit through a parallel interface.

The message operating system controls the timing and content of the terminal's messages using the variable parameters loaded from the 5110. The terminal's measured parameters and detected errors are stored in its input queue to be transferred to 5110 for analysis and correlation.

It is important to realize that the message operating system is used to load the two message units and the one bus loading unit with control information and data to be transmitted on the data bus. When this function is complete, the MOS issues a Run command to all message units and bus loading units, and from this point on these message units and bus loading units all independently compete to transmit their messages on the bus. After bus activity ceases, the MOS interrupts by reading transferred data within these message units.

4.6.3.2 Message Units

The message units simulate the external data rate of various classes of avionic equipment. The message units and bus loading unit act as a free-running, distributed-protocol system, with data being transferred at different rates between the units. The message unit terminal transmits data onto the fiber-optic bus when messages are ready to be transferred and when a quiet time proportional to the message unit's priority ID is detected. Each message unit terminal receiver accepts data when the sink code in the first word of the message matches the receiver's priority ID.

4.6.3.3 Bus Loading Unit

The bus loading unit transmits and receives data on the bus like a message unit. The bus loading unit serves as the means for varying the amount of bus loading. It alters its priority ID number for every bus request to contend with the message units for control of the data bus.

4.6.3.4 Detailed Hardware Description

The two message units and the bus loading unit consist of three major subsections, as shown in Figure 4-19. The major subsections are the bus operating system, the bus encoder/decoder, and the fiber-optic transceiver.

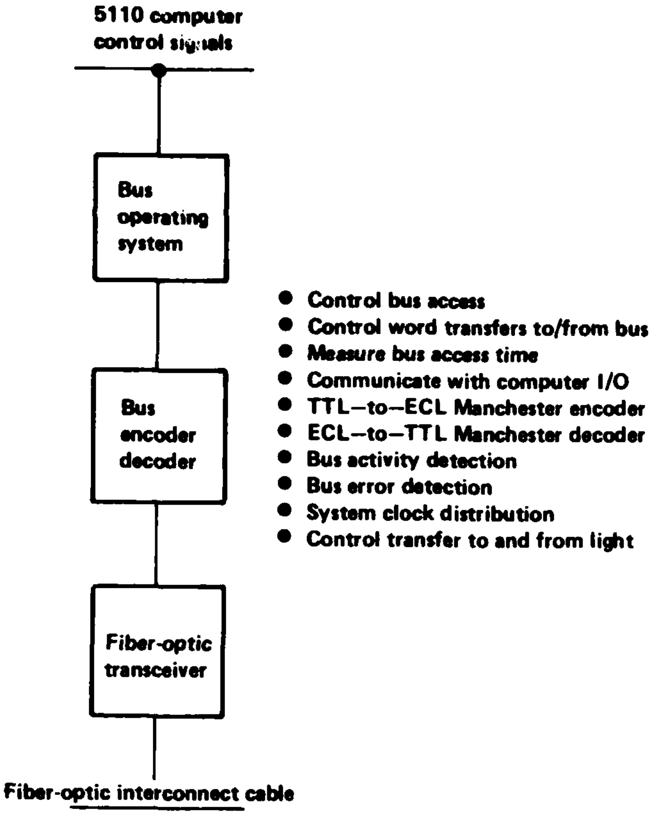


Figure 4-19. High-Speed Data Bus Message Unit

4.6.3.4.1 Bus Operating System

The bus operating system has the task of gaining control of the data bus and transferring data from its 1K-word output queue to one of the other message units. The bus operating system stores data received from other terminals. In one mode of operation, the time elapsed between a request and a bus access in gaining control of the data bus is measured by the bus operating system and stored in its 1K-word input queue.

4.6.3.4.2 Bus Encoder/Decoder

The bus encoder/decoder uses Manchester encoding and decoding techniques to convert a parallel TTL interface used by the bus operating system to a serial ECL interface used by the fiber-optic transceiver. Other functions of the bus encoder/decoder are sync detection, error detection, activity detection, and clock distribution for the message unit.

4.6.3.4.3 Fiber-Optic Transceiver

The fiber-optic transceiver controls the conversion between electrical signals and light.

4.6.4 SUPPORTING SOFTWARE REQUIREMENTS

The software needed to perform the system level test is divided into three sections:

- Message unit control modes
- Parameter set up of data, transmission rates, message lengths, and priority IDs
- Measurement parameter collection and analysis.

4.6.5 TEST PROCEDURES

Because of the three levels of testing to be performed, the procedure is divided into three subsections.

4.6.5.1 Component Level Procedure

The fiber-optic source is tested by feeding a pulse of current through it and measuring the light output power and rise time. The detector is tested by feeding a pulse of light into it and measuring the current output and rise time. The cable, connector, and multiport coupler are tested using the substitution method. A source/detector is connected without the component under test. Measurements are taken; then, the component is added and the measurements are taken again. The two measurements are compared to determine the effect of the component.

4.6.5.2 Subsystem Level Procedure

The data bus subsystem is tested by feeding a Manchester-encoded data signal into the data bus driver and measuring the output of the receiver. The ability of the receiver to reconstruct the original signal is verified in this manner.

4.6.5.3 Laboratory System Level Procedure

After the fiber-optic interconnect system is connected to all of the equipment in the laboratory, the system level testing begins. The special bus test then operates to determine the performance of the transmitter and receiver in a system environment. The operational bus test is then run to measure the system parameters needed to construct a fiber-optic, 50-MHz data bus.

Section 5
TASK IIa DETAILS

This section details the logic design, electro-optic design, software design, and final hardware of the proposed future avionic data bus demonstration system. Also reported are the results of tests performed in accordance with the Task IIa Development and Test Plan. The bus performance design goals are repeated here in table 5-1.

Table 5-1. Optical Bus Performance Design Goals

• Bit rate (serial)	50 MHz
• Number of terminals	16 minimum
• Maximum terminal separation	100 m
• Bit error rate (undetected)	1×10^{-12}
• Incomplete message rate	1×10^{-7}
• Timing	Asynchronous
• Protocol	Free access

5.1 DEMONSTRATION SYSTEM LOGIC DESIGN

5.1.1 INTRODUCTION

Herein is a detailed functional description of a 50-MHz fiber-optic data bus demonstration unit. Each of the three major logic subsystems (message operating system, bus operating system, and Manchester-encoder/decoder) is discussed. The system architecture is shown in Figure 5-1.

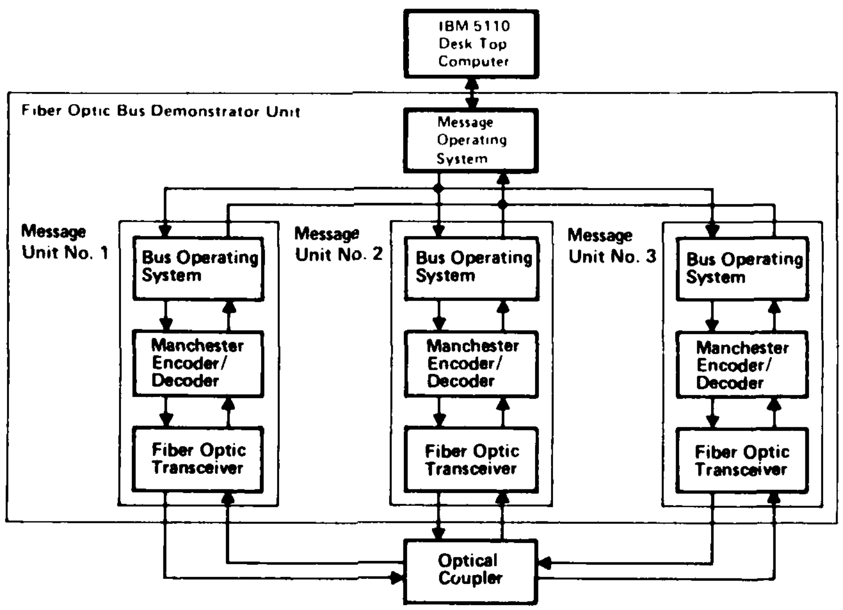


Figure 5-1. High-Speed Optical Data Bus Demonstration System

5.1.2 MESSAGE OPERATING SYSTEM

The message operating system (MOS) is a data formatter and translator that interfaces with the IBM 5110 computer and the bus operating system (BOS). The MOS includes a Z80-based controller with all memory and I/O interfaces contained on a single board. Features of this subsystem include an RS232-C-compatible serial I/O port; three 16-bit parallel TTL-compatible ports; up to 4K x 8 EPROM and 4K x 8 RAM; a tester interface for hardware/software debug; and buffering on all address, data, and control lines. Figure 5-2 is the MOS block diagram.

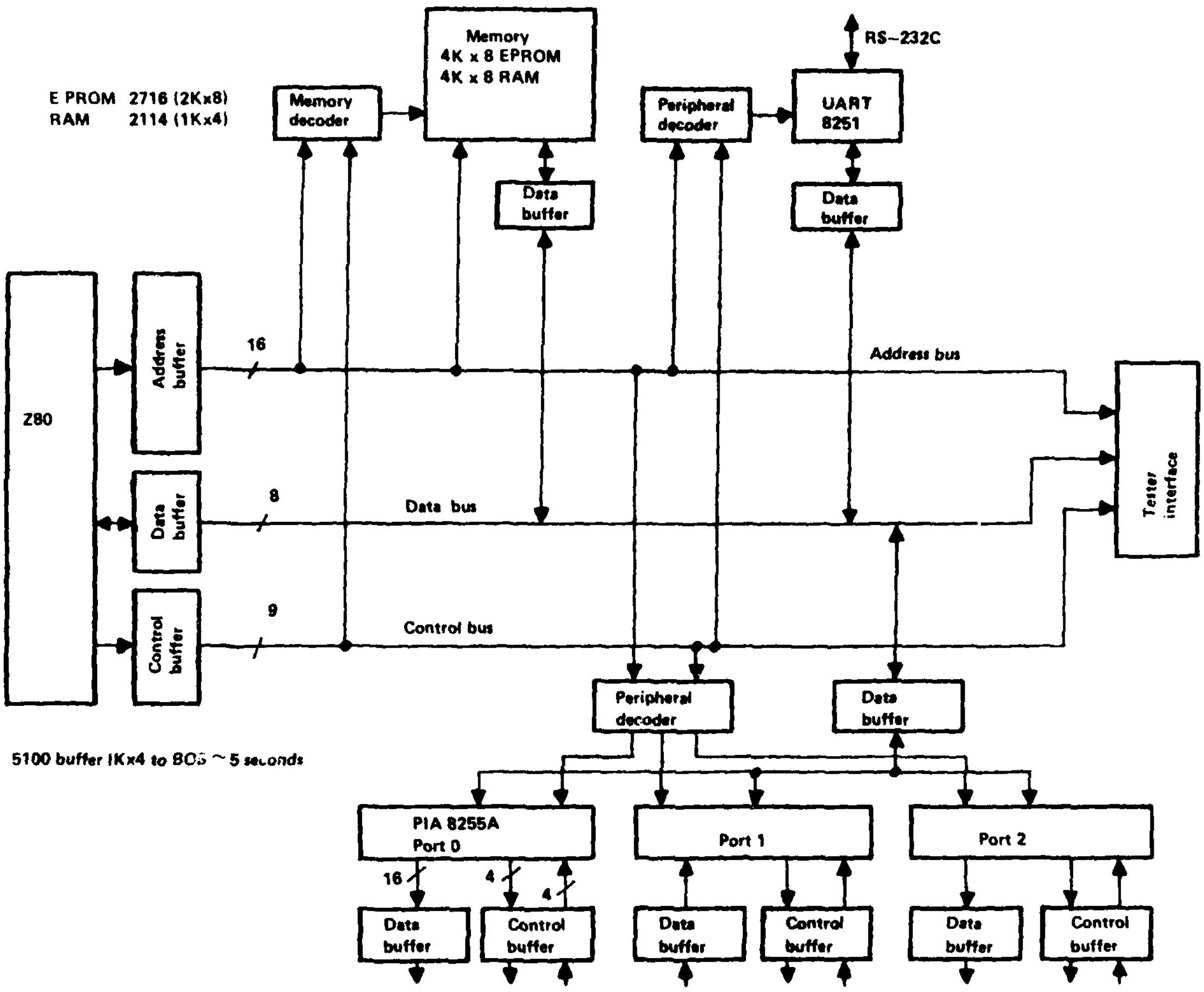


Figure 5-2. Message Operating System Block Diagram

5.1.2.1 Microprocessor

Within the MOS is a Z80 microprocessor. For simplicity, no interrupts or DMA capability are provided. All I/O is under program control. Only a minimum of support hardware is associated with the Z80; just that required to generate the control signals needed by memory, the I/O ports, and tester interface.

5.1.2.1.1 Tester Interface

All address and data lines are buffered by tristate drivers to supply the extra drive required by the I/O ports and the tester. This buffering allows the tester to monitor address and data buses and provides a means to gain control for debug purposes.

5.1.2.1.2 Memory and Decoding

Decoding is provided for 4K bytes of EPROM, 4K bytes of RAM, eight parallel I/O ports, two serial I/O ports, and a counter/timer. Features are added by populating the required sockets and reprogramming the EPROM as required. The EPROMs in this processor are Intel 2716 type 2K x 8. RAM memory is Intel 2114 type 1K x 4. RAM and EPROM are populated as needed.

5.1.2.1.3 Serial I/O

The serial I/O port communicates with the 5110 using standard RS-232-C levels (±12 V) and the standard 25-pin "D" shell connector. Data moving on this port is 7 bit ASCII, with even parity and two stop bits, at a data rate of 4800 baud. The data rate is selectable over the following range: 9600, 4800, 2400, 1200, 600, 300, 150, and 75 baud.

5.1.2.1.4 Parallel I/O

This subsystem supports three 16-bit parallel I/O ports that communicate with the MOS. These ports are dedicated to be either input or output and are programmed by jumper wires and software. Each port consists of 16 data bits and 4 control bits. Data transfers through these ports are based on a Data Ready/Acknowledge type protocol with two Data Ready's supported, one for control information and one for data. All inputs and outputs for each port are buffered by low-power Schottky tristate drivers. The ports provided are Data Output, Data Input, and Control Signal Output. Also provided are seven discrete inputs and five discrete outputs, which can be used as needed.

5.1.2.1.5 Self-Test

Each parallel output port can perform self-test (BIT). In the Self-Test mode, the port is disabled from transmitting or receiving data, and the BITE hardware is enabled. Various test patterns are then passed between the upper and lower 8 bits of the parallel port under software control. After all three ports have been tested, a system status word, a simple Go/No-Go indication, is output to the RS-232 port.

5.1.2.2 Data Transfers

5.1.2.2.1 5110-to-MOS Data Transfers

The MOS transfers data, commands, and status between the 5110 computer and the three message units. All transfers are initiated by the 5110. Commands to the MOS are contained in the first character of every message from the 5110. This first character is selected from the list of ASCII control characters. Any modifiers or data required for any command are immediately after the first character. Concluding every message is an End of Message character.

5.1.2.2.2 MOS-to-5110 Data Transfers

MOS-to-5110 transfers are similar in nature. The first character in every message is a control character that informs the 5110 what type of information is being transmitted: data or status. Any data transmitted is ASCII-encoded hex format.

Other routines included in the MOS software are Built-in Test (BIT), Load MOS Memory, and Execute Program in MOS Memory.

5.1.3 BUS OPERATING SYSTEM

The bus operating system (BOS) serves as the message unit/bus loader control element. The BOS consists of five basic subelements: the MOS interface logic, an output queue, an input queue, bus access control logic, and bus access time measurement logic. Figure 5-3 is a block diagram of the bus operating system.

5.1.3.1 MOS Interface Logic

The interface between the MOS and BOS consists of two parallel data buses and a set of discrete status bits. The MOS sends commands and data to the BOS on the "Output" bus and the MOS reads data from the BOS on the "Input" bus. The two data paths are common to all three BOS units with an expansion capability of up to 16 BOS units. The Discrete Status bits consist of one unique bit from each of the possible 16 BOS units.

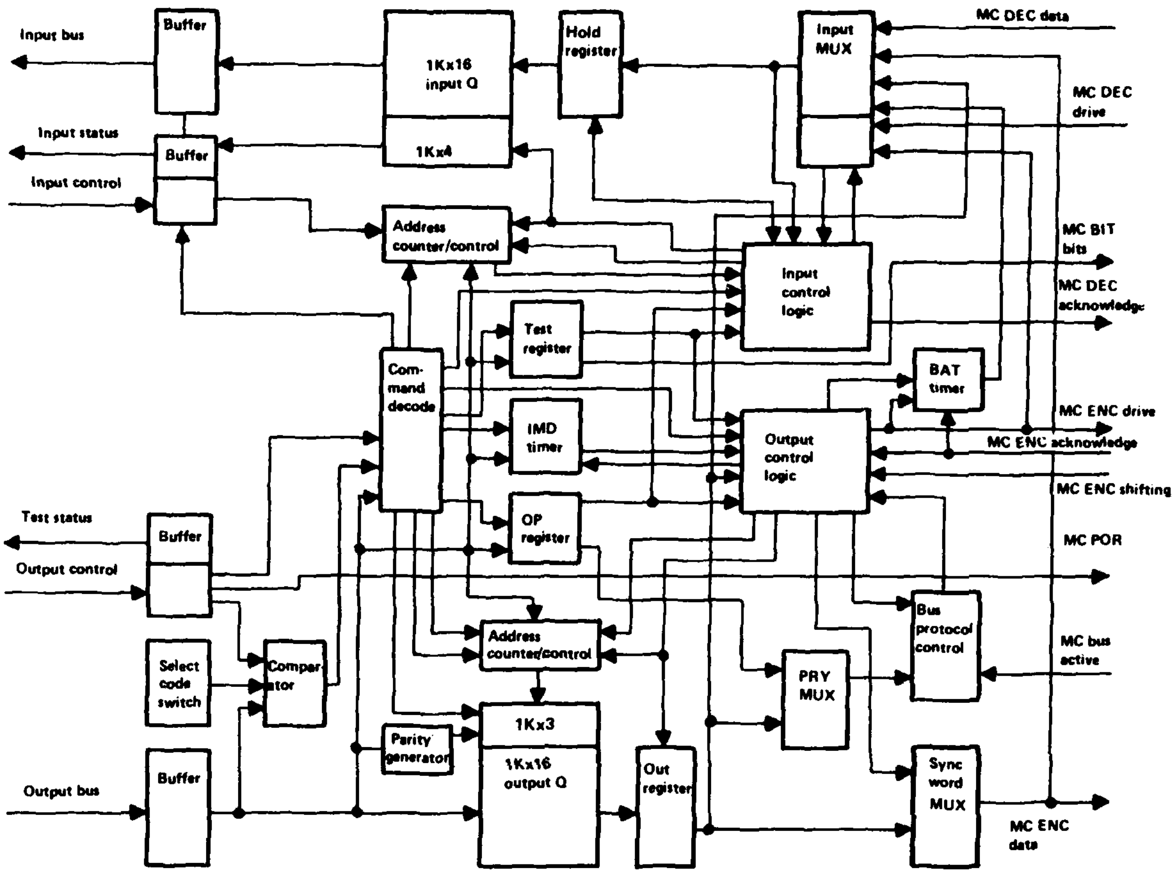


Figure 5-3. Bus Operating System Block Diagram

5.1.3.1.1 Output Bus

The output bus consists of 16 data bits, one "Command" Data Ready signal, one "Data" Data Ready signal, and one Power On Reset (POR) signal. The MOS issues all commands and all associated data words on the output bus. The presence of a command or data word is indicated by issuing the appropriate Data Ready. A BOS unit accepts or rejects commands and associated data as determined by the destination field in each command word. Figure 5-4 shows the command word format.

Destination					Not Used							Command			
15	14	13	12	11	10	9	8	7	6	5	4	3	2	1	0

Destination field

15	14	13	12

This 4-bit field is compared with a 4-bit message unit select code that is manually set via a DIP switch in each BOS unit. If the fields agree, the command is accepted.

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If this bit is a "1", the command will be accepted by all of the BOS units

Command field

3	2	1	0

0000	NOOP
0001	Load Operation Register (LD OP REG)
0010	Load Test Mode Register (LD TEST REG)
0011	Load Inter Message Relay Register (LD IMD REG)
0100	Load Out Queue Address Counter (LD OUT Q ADR)
0101	Load Out Queue Start of Message Word (LD SOM)
0110	Load Out Queue Word (LD WRD)
0111	Load Out End of Queue Word (LD EOQ)
1000	Load In Queue Address Counter (LD IN Q ADR)
1001	Read In Queue (RD)
1010	Clear Out Queue Address Counter (CLR OUT ADR)
1011	Clear In Queue Address Counter (CLR IN ADR)
1100	Not used
1101	Not used
1110	Not used
1111	Run

Figure 5-4. Command Word Format

5.1.3.1.2 Input Bus

The input bus consists of 16 data bits, one Data Valid bit, one parity bit, one Start of Message (SOM) bit, one End of Queue (EOQ) bit, and one Data Acknowledge signal. The input bus for a specific BOS is enabled via a Read In Queue command. A Read command must never be issued to more than one BOS at a time. Such a situation would cause a conflict on the input bus.

The MOS controls the operation of the input bus via the Read in Queue command and the Data Acknowledge signal. The MOS must first set the input queue address counter to the desired address (generally 000). It then issues a Read command, which enables the BOS input queue tristate drivers. The MOS then reads the 16 data bits as well as the 4 status bits. After the MOS has read the data,

it informs the BOS by issuing a Data Acknowledge, which increments the input queue address counter. The read sequence continues until the MOS detects an active EOQ status bit, which indicates the last meaningful word in the input queue.

5.1.3.1.3 Status Bits

In addition to the two bus links, each BOS unit has one discrete signal indicating that test operation has been completed. The MOS can monitor each BOS Test Complete status bit to determine when all the BOS units have completed their respective tests. Thus, the MOS can determine when it can issue a new command to the BOS units.

5.1.3.2 Output Queue

The output queue provides a large block of data that can be quickly accessed for high-speed data transmission on the fiber-optic (FO) bus.

The output queue is lK x 19 bits. It consists of 16 data bits, 1 parity bit, 1 SOM bit, and 1 EOQ bit. The output queue is loaded via three Load Out Queue commands. The SOM and EOQ bits are loaded as "1s" or "0s" depending on which load command is used. The parity bit is generated by the BOS.

The output queue must be loaded under the following rules. The start of the first message must be loaded at address 000. All messages must be at least two words long. All messages must conform to the word formats shown in figure 5-5.

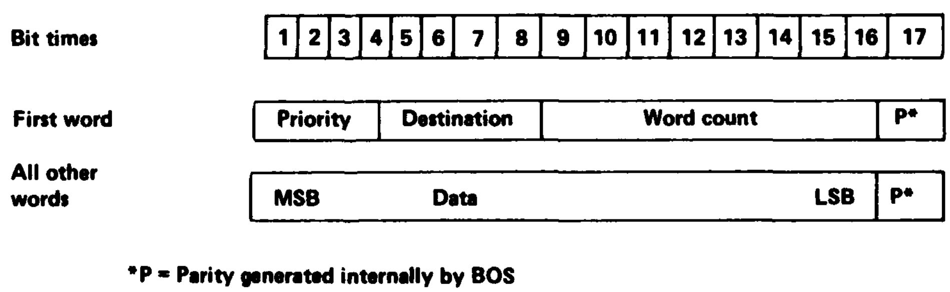


Figure 5-5. Word Formats

The contents of the output queue are transmitted to the Manchester encoder under the control of the bus access control logic. Transmission begins when a Run command is issued by the MOS. The 16 data bits and the parity bit are sent to the Manchester encoder for transmission. The SOM and EOQ bits serve as control bits for the BOS bus access control logic and are not sent to the Manchester encoder.

5.1.3.3 Input Queue

The input queue has a two-fold purpose. In one mode of operation, it stores data read from the FO bus. In another mode, the input queue stores bus access time information generated by the bus access time measurement logic.

The input queue is 1K x 20 bits, consisting of 16 data bits, 1 parity bit, 1 Data Valid bit, 1 SOM bit, and 1 EOQ bit. The source of the 16 data bits and the Data Valid bit varies, depending on the mode of operation. The SOM and EOQ bits are generated internally and serve the same basic functions as in the output queue.

5.1.3.4 Bus Access Control Logic

The bus access control logic controls the way data stored in the output queue is transmitted to the Manchester encoder and ultimately to the FO bus. The bus access control logic comprises an intermessage delay (IMD) timer, a bus protocol control element, synchronization logic, and output control.

5.1.3.4.1 Intermessage Delay Timer

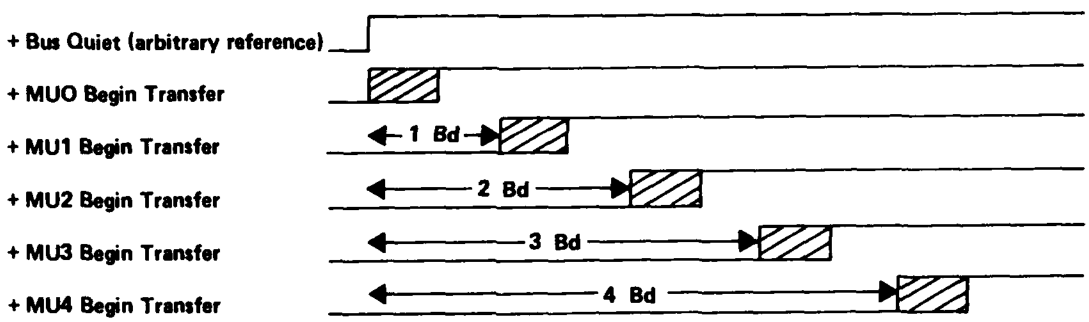
The IMD 16-bit interval timer provides a programmable time interval between the end of one message and the beginning of the next. The timer is started by loading it from the IMD register and enabling the count mode. The timer is clocked with a 12.5-MHz clock, producing a 80-ns LSB with a full-scale range slightly greater than 5.12 ms.

5.1.3.4.2 Bus Protocol Control Element

The bus protocol control element determines when a specific message unit can transmit a message on the bus. The control element employs a distributed protocol technique that assumes that no message unit is more than 50 m from the fiber-optic coupler. This implies that no two message units can be more than 100 m apart. There is no central bus controller to control bus transfers.

The bus protocol control element generates bus access "windows", which are the only times a message unit can initiate a message transfer on the bus. If a message is not available for transfer when a window occurs, the unit must wait for its next window. The bus protocol control element generates a bus window 1.25 μs times priority after a Bus Quiet condition is detected by the Manchester decoder.

If the Bus Quiet condition goes away (another unit has gotten on the bus) before the timeout occurs, the control element resets its logic and waits for another Bus Quiet condition. If a bus window does occur and the unit does not have a message to send, the second and all subsequent windows will come at 20-μs intervals. Any bus activity will force all units to reset their access logic to wait for their first window. Figure 5-6 illustrates the bus access technique.



Notes:

1. The maximum distance between a message unit and the optics coupler is 50 m. Therefore, the maximum bus length equals 100 m.
2. The maximum one-way propagation delay time (Td) is 537 ns.
3. All measurements are made with respect to the arbitrary reference shown

Time Delay Calculations:

Td = The one-way delay between the most widely separated terminals.

$$Td = \frac{L\,n}{3\times10^{8}\,m/s}$$, where L = separation length in meters
n = *index of refraction of the fiber* (1.61)
3×10^{8} m/s = speed of light in free space

$$Td = \frac{(100\ m)\ (1.61)}{3\times10^{8}\ m/s}$$

Td = 537 ns

Therefore, one roundtrip propagation time on the bus (Bd) will be 2xTd = 1.074 μs

Figure 5-6. Bus Access Technique

The bus protocol control element uses the same technique in all message units, with only the unit priorities being different. The bus protocol control element of a BOS programmed to operate as a bus loader operates slightly differently. The bus loader protocol element strips a 4-bit priority code from every SOM word and uses that field to determine its bus windows. Thus, a bus loader can simulate bus contention by more than one unit.

5.1.3.4.3 Synchronization Logic

The decentralized nature of the bus protocol technique causes a synchronization problem between the different units on the bus. There is no common clock between message units, and yet synchronization is essential to prevent bus conflicts. Synchronization occurs only when a message is transferred on the bus. If there is no bus activity from any unit, the internal bus access timers slowly "walk away" from each other and lose synchronization. To resolve this problem, the BOS unit with assigned priority 0 (highest priority) outputs one sync word if after 16 of its windows have occurred, there has been no activity and it does not have a message to send. The sync word has a destination field of 0, but the input queue will not store the sync word.

5.1.3.4.4 Output Control Logic

The output control logic handles data acquisition from the output queue and all handshaking with the Manchester encoder. The output control logic begins when a Run command is received. The first SOM word is read from the queue and stored in the output queue buffer register. The SOM word is held in the buffer register until the IMD timer has timed out and a bus access window is present. When this occurs, a Data Ready is issued to the Manchester encoder. As the Manchester encoder is shifting the first word on to the FO bus, the output control logic reads the next word of the message and issues another Data Ready at the appropriate time. This process continues until a new SOM word is detected.

When the new SOM is detected, the process stops and the IMD timer is restarted. The control logic then enters a Wait state until the IMD timer is restarted and a bus window again allows a message to be sent on the bus. This process continues until an output EOQ bit is encountered. When an EOQ bit is encountered, the control logic either stops outputting messages or it loops back to the beginning of the queue and continues, depending on how the BOS is programmed to operate.

5.1.3.5 Bus Access Time Measurement Logic

The bus access time (BAT) measurement logic consists basically of a 16-bit timer with a 40-ns LSB, yielding a full-scale measurement range of just over 5.12 ms. The timer measures the BAT of each message transmitted, permitting an accurate analysis of the bus efficiency.

The BAT timer is cleared and started when a Run command is received by the BOS. The contents of the counter are latched when the first Data Ready is issued to the Manchester encoder. When the BOS is programmed to do so, the first word of each message and its associated BAT data are stored in the input queue, as discussed in 5.1.3.3. In general, the BAT timer is cleared when the Manchester encoder indicates that it is shifting data out on the bus. When the encoder has finished shifting an entire message, it indicates this, and the BAT timer is restarted.

When running in the BAT test mode, the output queue continues to loop through its stored messages until the input queue indicates it is full, and the test is terminated.

5.1.4 MANCHESTER ENCODER/DECODER

The bus encoder/decoder uses high-speed Manchester encoding and decoding techniques to convert a parallel TTL interface used by the BOS into a serial ECL interface used by the fiber-optic transceiver. Figure 5-7 is a detailed block diagram of the Manchester encoder/decoder.

5.1.4.1 Manchester Decoder (MD)

The MD decodes a 50-MHz Manchester serial 20-bit data stream. The 20 bits consist of 3 sync bits, 16 data bits, and 1 parity bit, with each bit period equal to 20 ns. Figure 5-8 shows typical data word pattern as it would be seen at the MD input.

5.1.4.1.1 Fiber-Optic Decoder Interface

The data is received serially through a 10K ECL differential receiver (F10116) from the fiber-optic receiver.

5.1.4.1.2 Decoding Technique

The MD uses a four-stage trial-and-error decoder technique because of the high data bit rate (50 MHz) and because no two terminals are in sync.

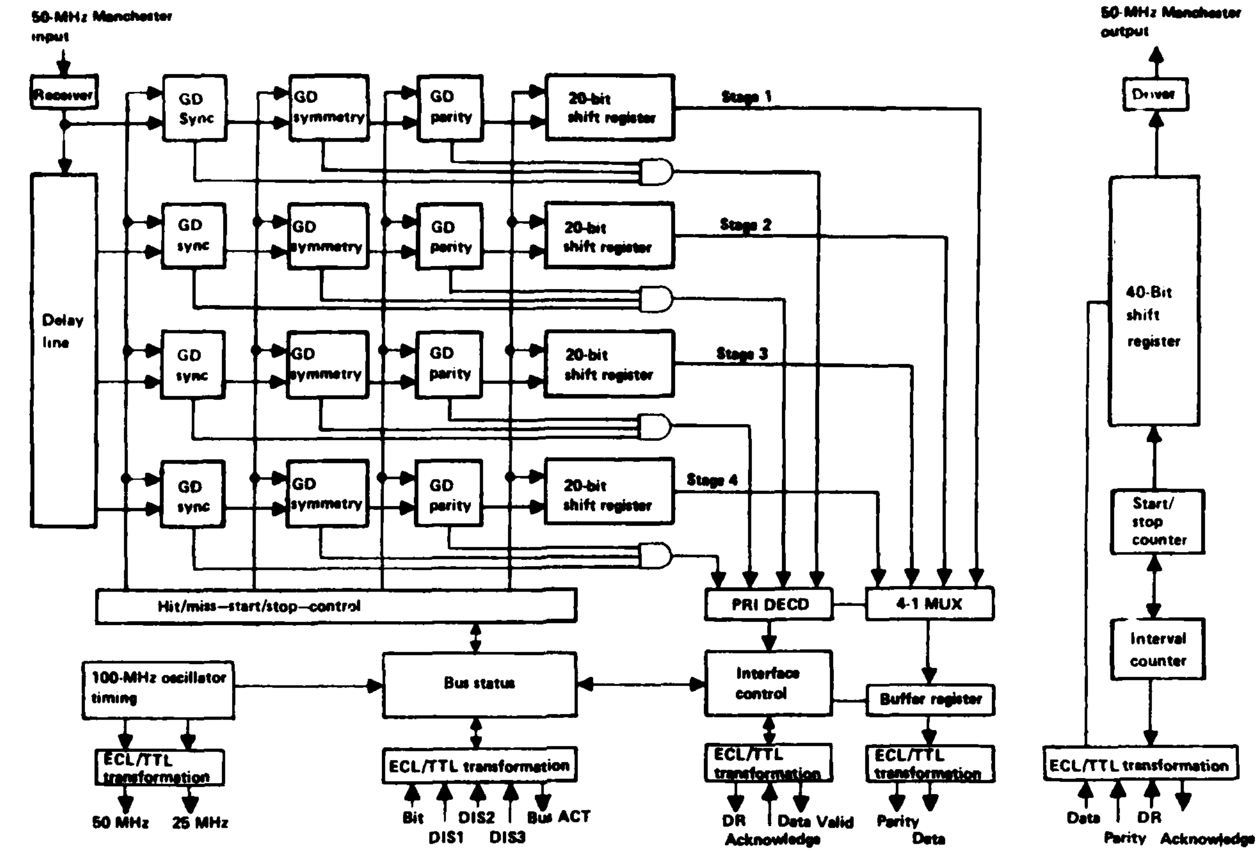


Figure 5-7. Manchester Encoder/Decoder Block Diagram

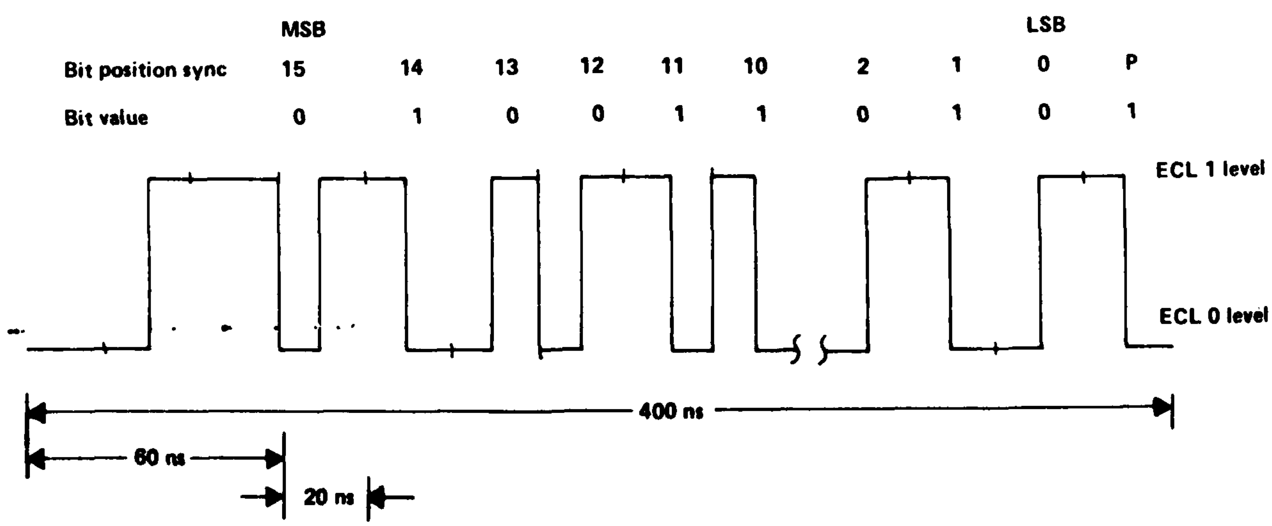


Figure 5-8. Typical Data Word Pattern at MD Input

The four MD stages strobe different phases of serial data into separate registers at a 50-MHz bit rate. Each data stream (word) is checked for good parity. Each bit field is sampled with a 100-MHz clock in an attempt to read the first and second half of the 20-ns bit field. The first 10 ns window should contain the true logic value, and the second, the compliment logic value if the data and clock are properly synchronized.

5.1.4.1.3 Decoder/BOS Interface

A priority decoder selects the data from the first register that indicates it has good data, as determined by the data sync, symmetry, and parity checks. The good data must be selected, multiplexed, and strobed into a buffer register before another 3-bit sync field is detected (< 60 ns). The data is then translated from ECL to TTL, and a Data Ready (DR) signal is issued to the bus operating system (BOS). The BOS must sample the 16 data lines, parity, and Data Valid, and acknowledge (ACK) the MD before another Manchester word is decoded and loaded into the buffer register (<400 ns).

5.1.4.2 Manchester Encoder (ME)

The ME encodes a 50-MHz Manchester serial 20-bit data stream. The data word is described in 5.1.4.1.

5.1.4.2.1 BOS/Encoder Interface

The bus operating system presents a Data Ready signal with 16 bits of data and 1 bit of parity on the interface. When BOS data is available, the ME translates this data from TTL to ECL. This data is then loaded into an output shift register when this register indicates an empty condition. Upon accepting this data, the ME issues an 'ACK' signal to the BOS.

5.1.4.2.2 Encoding Technique

The translated true and compliment data is loaded in parallel into a 40-bit shift register with the proper Manchester-encoded data parity and sync fields. The output shift register is then clocked at a 100-MHz rate.

5.1.4.2.3 Encoder/FO Interface

The data is serially shifted through a 10K ECL differential driver (F10192) to the fiber-optic driver.

5.1.4.3 Miscellaneous

5.1.4.3.1 Clock

The ECL Logic contains a 100-MHz oscillator and some countdown logic that produces 50-MHz and 25-MHz clocks, which are translated from ECL to TTL and sent to the BOS for timing purposes.

5.1.4.3.2 BITE

The BOS can program a direct wrap from the ME to the MD via a discrete line for BITE purposes.

5.1.4.3.3 Bus Activity

The MD continually monitors all bus activity and indicates to the BOS, through a discrete line, when there is dead time.

5.1.4.3.4 Discretes

The BOS provides three discrete disable lines to disable one or more of the decoder stages. These lines evaluate the Manchester decoder reliability and enables alternative Manchester decoding techniques.

5.2 HIGH-SPEED BUS ELECTRO-OPTIC DESIGN

5.2.1 FIBER-OPTIC DESIGN

The fiber-optic design begins at the transmitter circuit, which accepts data from the encoder and converts this electrical signal to light, using either a light-emitting diode (LED) or laser diode. The light is then transferred from these devices to a fiber-optic cable, which carries the light through connectors to an optical power divider. The latter device, commonly called an optical coupler, divides the optical power equally for distribution to each port on the bus via fiber-optic cable through connectors to a photodiode at each port. The photodiode converts the optical power into electrical signals, which are amplified by a receiver circuit at the photodiode to the electrical logic signal levels required by the system. The receiver sends its signals to the decoder.

5.2.1.1 Optical System Design

The system can serve 16 optical bus ports and therefore must have a 16-port optical coupler. To demonstrate a data bus, however, only three ports are required; therefore, only three of the 16 ports will be activated. Because the bus is being developed to fill avionic future requirements, a transmission line 100 m long has been

chosen, making the high-loss, bundle-fiber cable used in past systems inappropriate. Instead, single-fiber, medium-loss cables are used. For modularity, flexibility, and maintenance, a capability is provided for at least four connectors in any path between units on the bus. The input and output of the system must be a 50-Mb/s Manchester-encoded ECL signal.

Based on these expanded requirements, the optical system then requires the following:

- One 16-port optical coupler
- Three transmitter circuits
- Three LEDs or lasers
- Three receiver circuits
- Three photodiodes
- Ten cable connectors (minimum)
- Three hundred m of optical cable.

Given these components, a baseline optical loss budget for the optical system can be assigned, reflecting the expected system component contributions. The total loss sets the design points for the transmitter and receiver circuits. The baseline* optical loss budget follows:

Item	Loss (dB)
A 16-port coupler	16
Four connectors @ 1.5 dB	6
One hundred m of cable @ 10 dB/km	1
Output coupling loss	1

5.2.1.2 Optical Component Design

Because this developmental project was concerned with state-of-the-art components projected for use in the 1985-1990 time frame, manufacturers of optical components were surveyed. Over 125 companies were contacted and visits made to those which had an optical component fitting the task. Each vendor visited was asked to formally quote price and delivery of any system component within their capability. Technical discussions with engineering and manufacturing people yielded a free exchange of nonproprietary information at each company visited, which, when summarized, gave a reasonable estimate of the state-of-the-art of each system component. The following are conclusions reached from the survey.

LEDs have a definite rise time capability limit as well as a defined power output limit. Both are somewhat controlled by current density, but are controlled by device material and structure to a greater extent; the result being a tradeoff between either speed or power, similar to that of gain/bandwidth in an amplifier.

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*Actual measurements on delivered components gave a loss of 27 dB total

The efficiency of LEDs remains relatively constant until a speed of about 20 ns is introduced into the design. At this point, efficiency must be traded for speed. At about 10 ns rise time, the "knee" of the curve is reached, and further increases in speed cause greater penalties.

The highest speed claimed by any manufacturer was 3 ns, and at this speed, less than 5% of the normal LED efficiency would result. All manufacturers agreed that a 2-ns LED appeared to be unattainable and extremely questionable for development in the 1985-1990 time frame. Due to the simplicity of LED driver circuits, however, LEDs are of interest in this program, as is the extension of capability of state-of-the-art devices.

5.2.1.2 Lasers

Lasers definitely have the power and speed required for this program. Power outputs in excess of 2 mW and speeds less than 1 ns are routinely achieved within the industry. There are, however, other considerations when utilizing solid-state lasers: temperature dependence of operation, complex control circuitry, and safety. To use a laser in a military electronics environment, voltage control, temperature control, and optical feedback are required. In addition, these devices are Class III b category according to Bureau of Radiological Health Regulatory documents and are damaging to eyes and skin at close distances.

Almost all companies involved with laser manufacture have extensive development programs aimed at the communications industry and fiber optics in particular. It is in these devices that manufacturers expect the greatest improvements for the 1985-1990 time period.

5.2.1.2 Optical Couplers

Only four companies have fabricated optical couplers in the U.S.; three of these were visited. These three agree that single-fiber couplers are both practical and desirable over bundle-fiber couplers. The favored design is the transmissive-type coupler, because interface losses are easier to control. However, both intrinsic loss and port-to-port variations will likely be greater in single-fiber couplers than in bundle-fiber couplers. The state-of-the-art in bundled couplers produces intrinsic losses of about 5 dB. Therefore, a loss of 6 dB or greater will likely occur in single-fiber couplers.

5.2.1.2.4 Cable

It had been hoped to find a manufacturer that produced a plastic-clad, fused-silica fiber meeting the military temperature range and which could be properly terminated. This was not the case. One manufacturer claimed it could produce such a cable, but could not prove successfully manufacturing such a cable. There was no other choice

but to choose a glass-clad, glass-fiber structure and use the largest available size to minimize connector interface and input interface losses.

5.2.1.2.5 Connectors

Many connector manufacturers are actively pursuing the design of single-fiber connectors, a state-of-the-art design. Prices typically exceed 100 dollars per connector; therefore, the optimum connector will not be sought out; rather, a cost-effective choice that reasonably controls connector interface losses.

5.2.1.2.6 Photodiodes

There are still two basic choices in photodiodes: (1) PIN structure diodes, and (2) avalanche diodes. Complex control circuits and high voltage (over 150 V) are still required for avalanche detectors, which have a 10-dB advantage over PIN diodes. However, PIN photodiodes are available from more manufacturers, and design improvements have been made in areas of low noise, fast speed, small size, and lower voltage operation. PIN photodiodes are a clear choice for this application.

5.2.1.2.7 Component Specifications

The baseline optical budget, the industry survey, and manufacturers replies to formal requests for quotes led to the following component procurement specifications:

- Laser Diode - Order per General Optronics Corp. Quote 2076, 18 May 1979. Description: Model GOLT-3 laser transmitter, with a substitution of Gallite 3000 LC-MS fiber cable type, 0.5 m long, terminated with Amphenol connector type 906-113-5000. Transmitter to be optimized for an ECL level input (-0.8 V to -1.6V) signal. The laser must conform with all applicable sections (parts 1000-1040) of the BRH regulations (210FR, Chapter 1, Subchapter 6, Radiological Health.)
- LED - Order per Spectronics, Inc. Quote Q 10771, 20 April 1979. Description: Item 4 plus 7; part number SE-0352-003, with part number SPX-4089-XXX attached. Light-emitting diode (820 nm) with T_r = 12 ns, which supplies 200 μW of power out of a mode stripped Gallite 3000 LC fiber. This device includes an attached 0.5 m Gallite 3000LC-MS fiber, which is mode stripped and terminated with an Amphenol 906-113-5000 connector or equivalent.
- LED - Order per Texas Instruments. Quote of 12 April 1979, Gene Dierschke. Description: Light-emitting diode (860 nm) with T_r = 5 ns typical (T_r = 6 ns max.) at 100 mA I_f. This device includes

a 0.5-m Gallite 3000LC-MS fiber pigtail terminated with an Amphenol 906-113-5000 connector attached to the device.

- Photodiode - Order per Spectronics Inc. Quote Q10771, 20 April 1979. Description: Item 5 plus 7, part number SD-0322-002, with part number SPX-4089-XXX attached. Silicon PIN diode, 100% T_r = 3 ns at 5 V, depletion voltage = 3.5 V. This device includes an attached 0.5 m Gallite 3000LC-MS fiber terminated with an Amphenol 906-113-5000 connector or equivalent.
- Fiber Optic Cable - Order per Galileo Electro-Optics Corp. Quote of 24 May 1979. Item 3000LC-MS10. Description: Fiber-optic cable type 3000LC-MS 10-m long. Terminated with Amphenol 906-113-5000 connectors on both ends.
- Fiber-Optic Connectors - Fiber-optic connector for Gallite 3000LC-MS fiber, 906-113-5000 - Fiber-optic connector adapter 905-120-5000.
- Optical Coupler - Order per Spectronics Inc. Quote Q107701, 20 April 1979. Description: Item 1, part number SPX-3613-322*,single-fiber, 32-port radial coupler. This is a transmissive coupler with 16 input ports and 16 output ports. The coupler is to be compatible with Gallite 3000LC fiber type (0.204-mm core) and Amphenol 906-113-5000 connectors or equivalent. The excess loss shall not exceed 5 dB, and the port-to-port uniformity must be within 3 dB. Nominal size, 5 x 5 x 1.5 inches.
- Optical Coupler - Order per Spectronics Inc. Quote Q10771, 20 April 1979. Description: Item 3, part number SPX-3613-322*, 10-m, single-fiber, 32-port radial coupler with 10-m pigtails of Gallite 3000LC-MS fibers terminated with Amphenol 906-113-5000 connectors or equivalent. This is a transmissive coupler with 16 input ports and 16 output ports. This coupler will have no connector at the fiber pigtail/coupler interface. The excess loss, including the fiber, shall not exceed 6 dB, and the port-to-port uniformity must be within 3 dB. Nominal size, 5 x 2 x 1 inches.

The laser diode chosen was an analog device; no digital models were available. This device required proper light level adjustment for digital use and an additional buffer circuit to isolate its analog input from digital and power noise.

The LEDs were ordered as the best representatives of state-of-the-art devices optimized for speed/power. Note that no other companies formally responded.

The photodiode ordered is a new device that will operate at a 5-V bias with a rise time of 3 ns. At 15 V, it will operate at 1 ns or less. Presently, this device is in a class by itself, and low-voltage operation is extremely desireable.

The fiber-optic cable is the largest diameter glass-clad fiber being manufactured in the U.S. Its core diameter is 204 μm.

The connectors represent a cost-effective choice. Connector loss is expected to be about 1.5 dB - the cost is $13.00. The

highest quality optical connectors available have losses that average between 0.5 and 1.25 dB. The price of these is from 10X to 20X higher than the choice made.

Two couplers were ordered. Both are 16-port transmissive types, but one utilizes 10-m pigtails, which should produce a harness type system with lower overall losses.

5.2.2 ELECTRICAL DESIGN

5.2.2.1 Laser Transmitter

The laser transmitter, a purchased device, is a model GOLT-3 manufactured by General Optronics Corp. This unit was adjusted at the factory to be compatible with ECL-level signals.

The laser transmitter, an analog device, is always biased on, even with a logic "0" input. This condition will allow any noise on the logic input to modulate the laser, injecting noise into the system. Further, the input is a single-ended 50-Ω input, and the system furnishes double-ended input signals to a 110-Ω load. A simple buffer circuit added to the laser eliminated both these problems. A schematic of this circuit is shown in Figure 5-9.

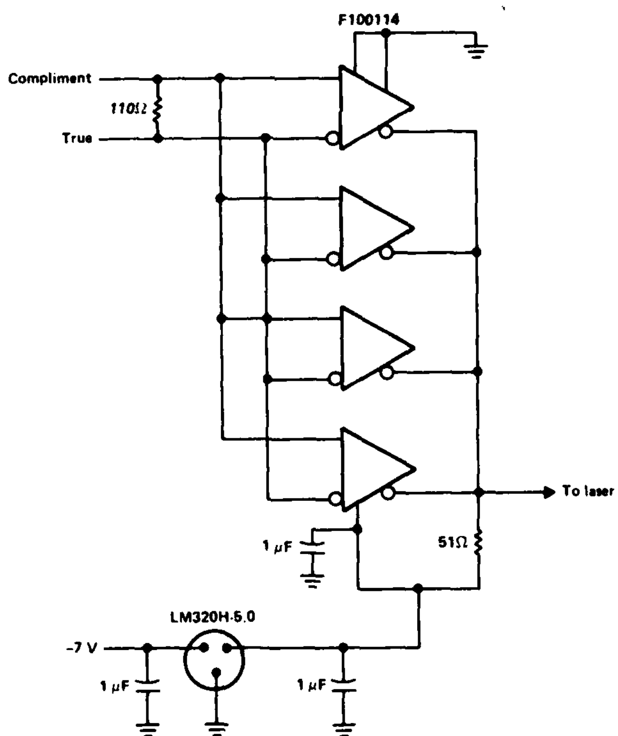


Figure 5-9. Laser Transmitter Buffer

The circuit is simply a double-ended line receiver (F100114) with all gates in the package paralleled to provide laser drive current. The input is terminated with the proper impedance (110Ω), and the output source impedance is terminated in 51Ω to -5.2 V. The circuit utilizes the same voltage supplied to the laser (-7 V) and regulates it to the voltage required by the gate (-5.0 V). This circuit utilizes capacitors at the input to the regulator, output of the regulator, and at the device package to ensure proper device decoupling and minimize any possible power-line-injected noise.

The laser and the buffer circuit are mounted on a single printed-circuit card, and covers were fabricated to shield the buffer from radiated noise. The total assembly then forms a buffered laser transmitter with a double-ended ECL input (true and compliment) and a -7-V and ground power supply.

5.2.2.2 LED Transmitter

Although not used in the final demonstration system, a LED transmitter was designed, built, and tested. The transmitter, Figure 5-10, consists of an input differential line receiver, a voltage translator, and an output current driver.

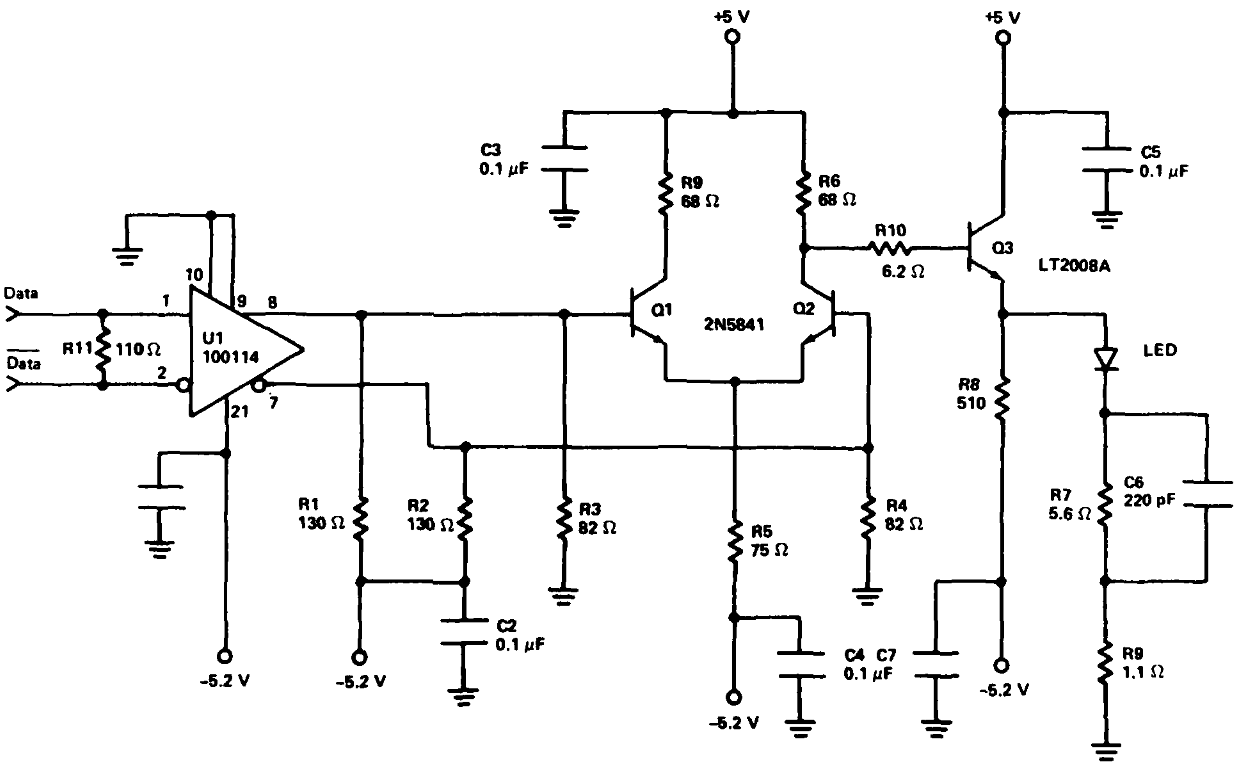


Figure 5-10. LED Transmitter

The input to the LED transmitter is a double-ended ECL signal from an ECL line driver. However, any double-ended signal with a common mode range between +0.1 V to -2.7 V and a differential voltage of at least 150 mV will be sufficient to drive the LED. Data is sent to the transmitter on twisted-pair wires, with the true data going to pin 1 and compliment data going to pin 2; this will result in light output for a logic "1" and no light output for a logic "0". The 110-Ω resistor terminates the line in its characteristic impedance.

The outputs of the line receiver are terminated in a Thevenin equivalent of 50Ω to -2 V and switch between the typical ECL levels of -0.955 V and -1.705 V. This line receiver has a typical propagation delay of 1.4 ns and transition time of 0.7 ns.

A current mode switch was used for the voltage translator because of its inherently fast switching times, which occur because the transistors are never saturated. Rather, the current is just steered from one transistor to the other.

The emitter resistor common to both transistors was chosen such that 37 mA will be switched from one 2N5841 to the other. With a collector resistor of 68Ω, the voltage on the collector of Q2 will swing from +2.5 V to +5 V. The 2N5841 transistors were used for Q1 and Q2 because they were designed to provide fast switching times in current mode circuits at collector currents up to 80 mA. At collector current around 37 mA, these transistors have delay times of less than 0.5 ns and rise and fall times of less than 0.3 ns.

An emitter follower drives the LED with a peak initial current of 400 mA and a final current of 250 mA. The transistor used as the emitter follower is a TRW LT2008A designed for high-speed, high-current pulsing applications and has an f_t of typically 3 GHz.

The LED is biased at a forward current of about 40 mA, which results in a forward voltage drop of about 1.5 V. The ideal current waveform for turning on and off the LED in the shortest time is shown in Figure 5-11.

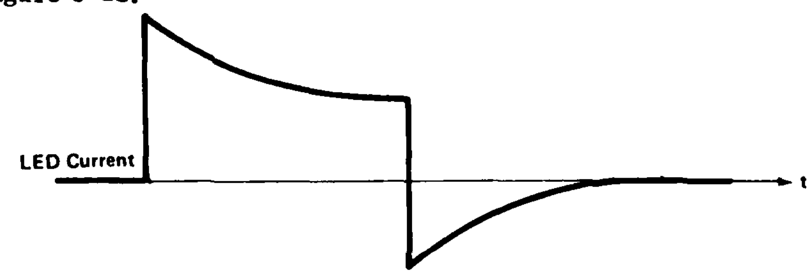


Figure 5-11. Ideal LED Current Waveform

The initial peak of current on the leading edge of the pulse builds up the minority carrier stored charge in the LED, giving the LED light output a faster rise time than if the current was just a step function. When the LED is turned off, the reverse current spike removes the stored charge due to minority carriers, resulting in a shorter fall

time than with a step function current waveform. This speed up was accomplished by placing a 220-pF capacitor in parallel with the current-limiting resistor, R7.

The actual waveform for the LED current is shown in Figure 5-12. The leading edge current peaks at 450 mA and settles to 250 mA at the end of the pulse. The trailing edge current spike is -400 mA. This gives a speed-up factor of 450 ÷ 250, or 1.8. Thus, according to Biard, the light rise time should be reduced by a factor of 3.2 over a LED current with a step response. The rise and fall times of the current were measured, using a scope and preamp with a 1.4-ns rise time, to be 1.8 ns. Thus, the current rise and fall times are about 1.1 ns.

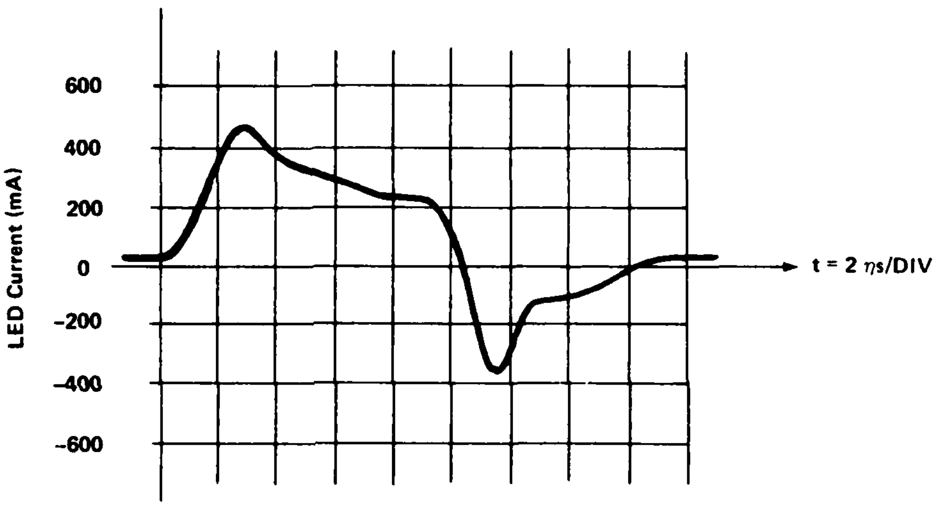


Figure 5-12. Actual LED Current

5.2.2.3 Receiver

Section 4.5.5 described a proposed receiver design approach. Its block diagram is repeated here in Figure 5-13.

The preamplifier is the cascoded bipolar transimpedance amplifier of Figure 5-14. The feedback resistor value of 3,600Ω was chosen as a compromise among gain, noise, and upper frequency response. Because in this system the principal source of noise is the receiver preamplifier itself, we must calculate this noise to determine the minimum input signal for maintaining a sufficient signal-to-noise ratio for achieving the required bit error rate (BER).

The equivalent input noise current for a bipolar transimpedance amplifier is

$$* \quad i_n = \sqrt{4kT\left\{\left(\frac{1}{R_f} + \frac{qI_e}{2kT\beta}\right) B_n + 64\, C^2 B_n^3 \;\; \frac{N}{M^2}\left(\frac{kT}{2qI_e} + r'_{bb}\right)\right\}}$$

*See Appendix B

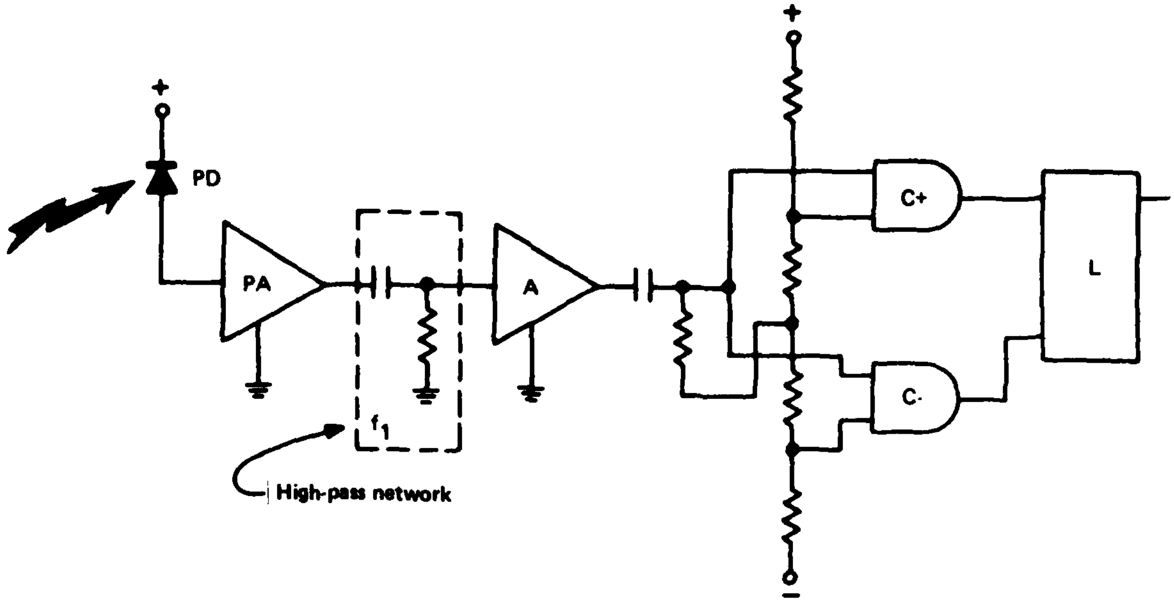


Figure 5-13. Receiver Block Diagram

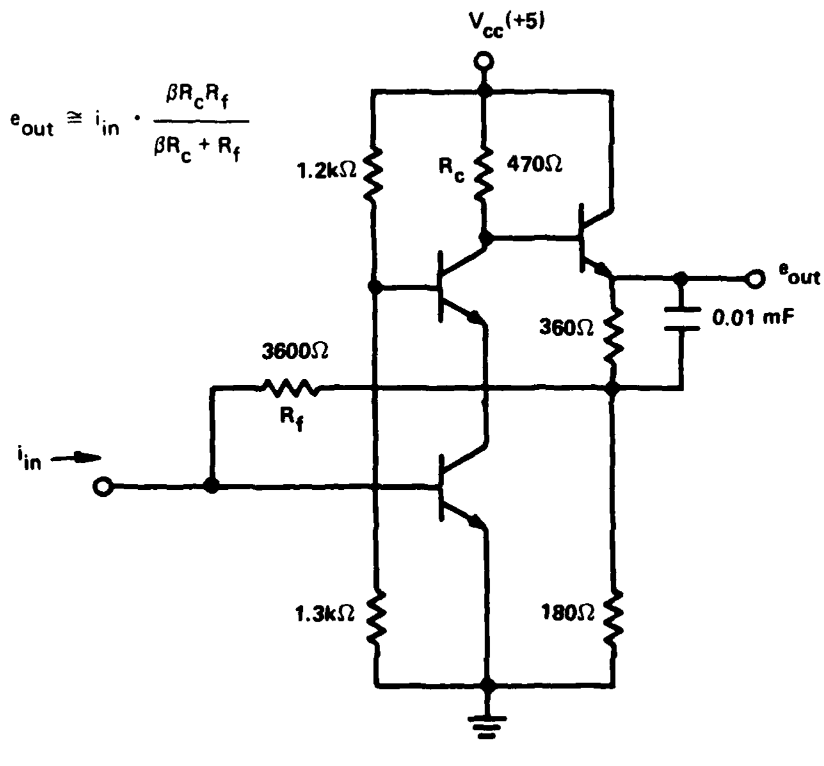


Figure 5-14. Cascode Transimpedance Preamplifier

where

R_f is the feedback resistor
k is Boltzmann's constant (1.38×10^{-23})
T is the Kelvin temperature
β is the transistor current gain
q is the electron charge (1.6×10^{-19})
I_e is the transistor emitter current
C is the circuit input capacitance
r'_{bb} is the transistor base resistance
B_n is the noise bandwidth

$$N = 1 + 2\alpha \qquad \alpha = \frac{f_L}{f_H}$$

$$M = \left(\frac{1-\alpha}{1-\alpha^2}\right)^2$$

B_n is defined as $\dfrac{\pi f_H M}{4}$

where

f_H is the amplifier high-frequency cutoff caused by two poles at f_H

f_L is the amplifier low-frequency cutoff caused by one pole at f_L

Using values for circuit elements chosen for the amplifier, one calculates an equivalent input noise current of $\approx 30 \times 10^{-9}$A. To achieve a BER of 1×10^{-12}, a signal-to-noise ratio of *7.03 is required.

If we now look at the requirements for the postamplifier, one needs to set the threshold voltage at least 7.03 times higher than the rms noise voltage at the comparator input to avoid false pulse detection at rates less than 1×10^{-12} for 50-MHz data rates. The calculated mid-frequency gain of the preamplifier, high-pass filter, and postamplifier up to the comparator results in an rms noise voltage at that point of 30 mV. Then, $7.03 \times 30 \cong 211$ mV for the threshold.

To obtain a probability of 1×10^{-12} for a missed pulse, we must exceed the threshold by an equal amount at signal peaks. Thus, starting with zero reference, the threshold is 211 mV, and the minimum peak signal is 422 mV for a BER of 1×10^{-12}.

The response of the receiver to a Manchester-encoded optical waveform is not the same as the mid-frequency gain of the receiver. The calculated response to such a waveform results in requiring a peak signal current into the transimpedance amplifier of = 1×10^{-6} A. Because this is the PIN photodiode current in response to a light input,

*From error function tables (erfc).

if it is divided by the photodiode responsivity, the required light power will result. The average measured responsivity of the photodiodes purchased was 0.24 A/W. This means 4.16 mW$(1 \times 10^{-6} + 0.24)$ is required at the photodiode to achieve the system goals. To characterize the receiver according to convention (light input required for a signal-to-noise ratio of one), the receiver sensitivity is -39 dBm, where 0 dBm is 1 mW.

Section 5.2.1.1 showed a possible 24-dB attenuation for a system starting with the input fiber. If 4.16×10^{-6} W of signal are needed at the photodiode, this translates to 1.04 mW into the fiber from a transmitter. If the actual 27-dB system is used, 2.08 mW minimum will be required. Actual laser transmitter measurements have shown a peak-to-peak output of about 4.5 mW from the pigtail. Passing through a 1-dB connector leaves 3.6 mW peak-to-peak; adequate to meet the requirements.

To handle the expected dynamic range, the postamplifier consists of four stages of limiting differential amplifiers. Figure 5-15 is a schematic of one of the identical stages. A current source to the emitters limits the total current that can all be in one side or the other. Because the change in current through the equivalent output collector resistor is limited to this total current, no matter what the input voltage from the preceding stage does, the output voltage is also limited to about 1.5 V peak-to-peak.

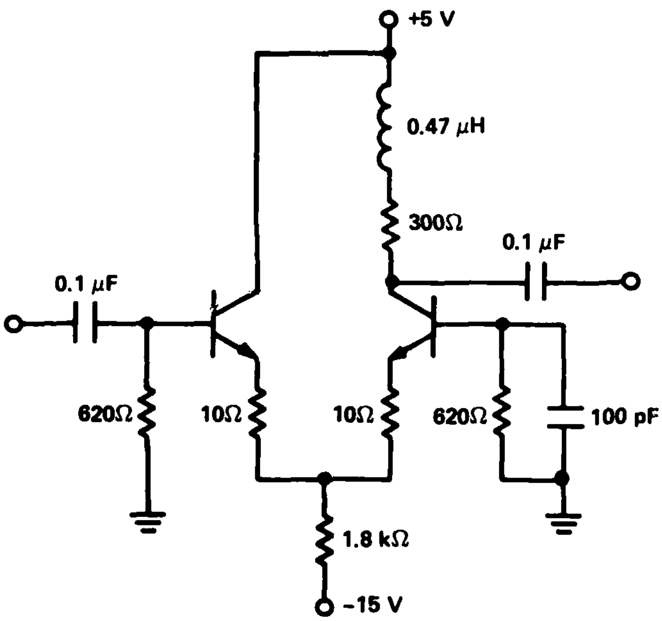


Figure 5-15. Limiting Differential Amplifier

Detailed results of BER measurements with various thresholds and system attenuations are shown later in Table 5-5, page 5-42. Figure 5-16 is the schematic of the total amplifier receiver.

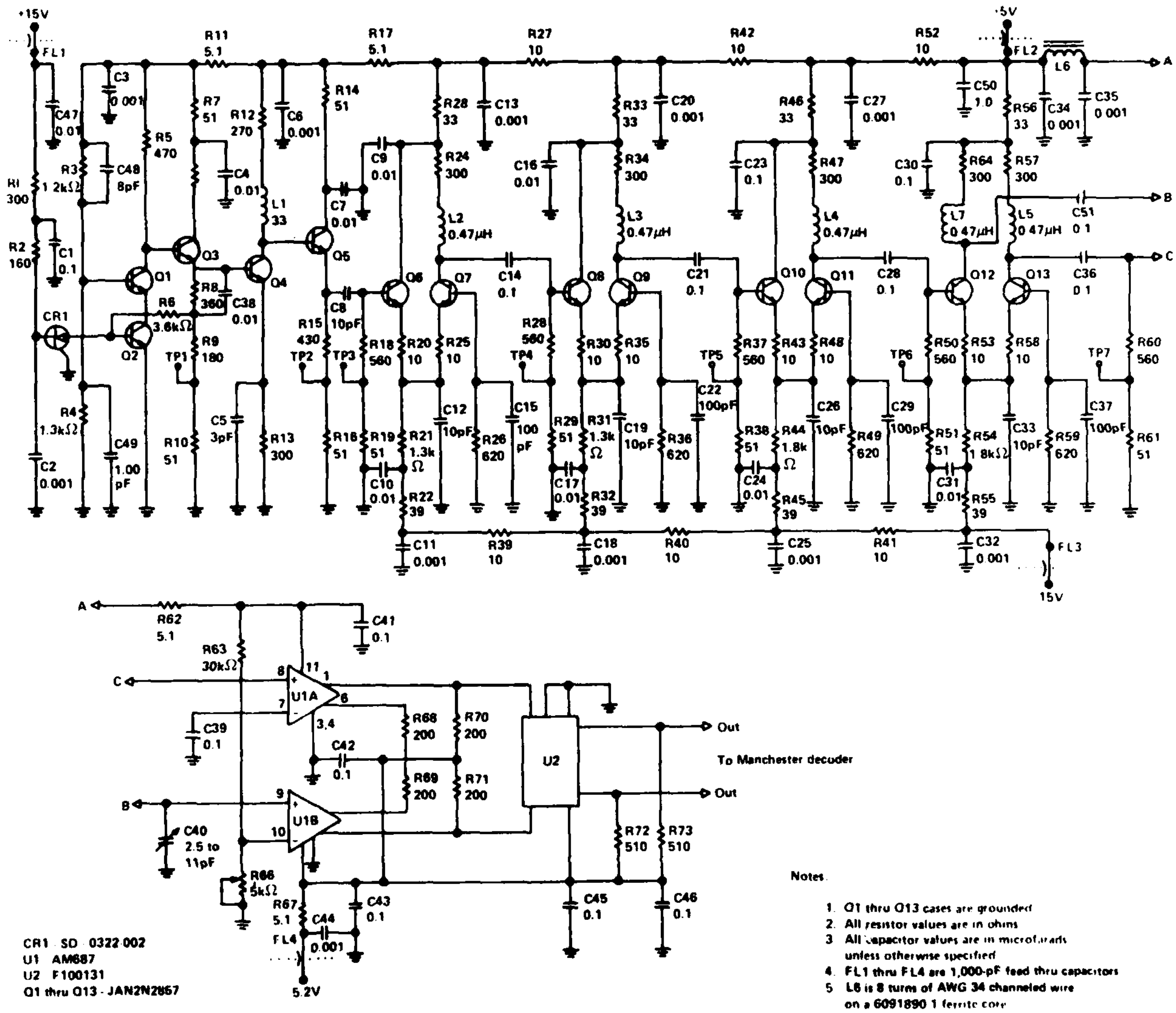


Figure 5-16. Receiver Schematic

5.3 DEMONSTRATION SYSTEM HARDWARE DESCRIPTION

Figure 5-17 is a photo of the high-speed fiber-optic demonstration system. Except for the IBM 5110 and the optical coupler, the entire system is contained in the 20 x 22 x 30-inch cabinet shown. The system is assembled on a sliding platform, allowing for easy removal from the cabinet.

The important feature of the packaging scheme is accessability. Each message unit logic section is mounted on its own hinged gate. The gates can be opened to allow easy access to the component side of all the boards and fiber-optic transmitters. The transmitters are located inside the supporting tower structure and are not readily visible until the gates are open. The fiber-optic receivers are located directly below their respective logic sections.

The power distribution for the demonstration system is divided into a logic and a fiber-optic power system, which are maintained independent of each other to ensure minimum noise on the supply voltages to the sensitive fiber-optic receiver subassemblies.

All logic functions are designed with commercial parts interconnected on wirewrap boards. This package technique produced the most flexible, cost-effective approach for this program. However, hardware must be redesigned using military-grade integrated circuits mounted on multilayered printed circuit (PC) boards for actual system applications.

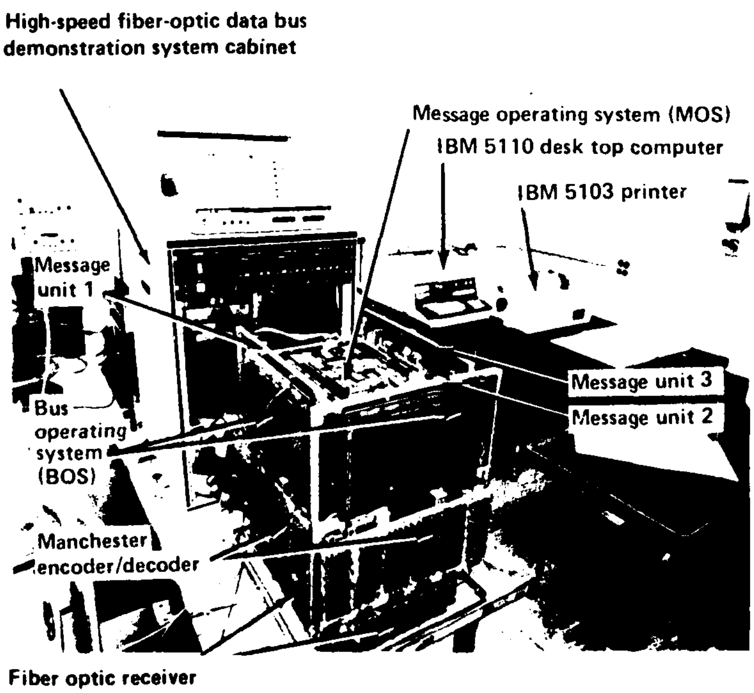


Figure 5-17. High-Speed Fiber-Optic Data Bus Demonstration System

5.3.1 MESSAGE OPERATING SYSTEM

The message operating system (MOS), in conjunction with the IBM 5110 desk top computer, is a simulation of avionic bus devices. In an operational system, the MOS would be replaced with separate pieces of avionic equipment tied together on a common system data bus.

The MOS is a microprocessor-based (Z80 μP) design requiring a certain amount of μP software, as explained in 5.5. Most all other functions are derived using large-scale integrated (LSI) transistor - transistor logic (TTL) devices wirewrapped on a common board.

5.3.2 BUS OPERATING SYSTEM

The bus operating system (BOS) functions include storing data for transmission on the fiber-optic (FO) bus, controlling bus access, and storing appropriate data received on the FO bus.

The BOS is designed using an even mix of LSI, MSI, and SSI TTL devices, all interconnected on a wirewrap board. Field programmable logic arrays (FPLA) provide the major hardware control sections of this design.

5.3.3 MANCHESTER ENCODER/DECODER

The Manchester encoder/decoder converts a two-directional parallel TTL interface from the BOS to a serial emitter-coupled logic (ECL) interface used by the fiber-optic transceiver.

The encoder/decoder is designed using a mix of MSI and SSI ECL devices. The 10K ECL family was used because it is the fastest logic family known that can be wirewrapped, allowing for a flexible design in this demonstration system. The design was packaged on a special wirewrap board that optimizes the characteristics of ECL devices. To ensure this design will function in a military environment, it must be redesigned using the 100K ECL family on multilayered PC boards.

5.4 DEMONSTRATION SYSTEM OPTICAL HARDWARE DESCRIPTION

5.4.1 OPTICAL RECEIVER

The optical receiver was packaged utilizing discrete devices mounted on a PC card. The basic structure was separated into compartments for shielding purposes. The mechanical design procedures were similar to those of an IF amplifier because the receiver is a high-gain, 50-MHz wideband amplifier. Figure 5-18 shows the receiver with the cover removed.

No attempt was made to minimize the size of the receiver because this program called for development application of projected system

needs. This package technique produced the most flexible, cost-effective approach for this program, but would most likely be a hybrid or monolithic package technique in actual application

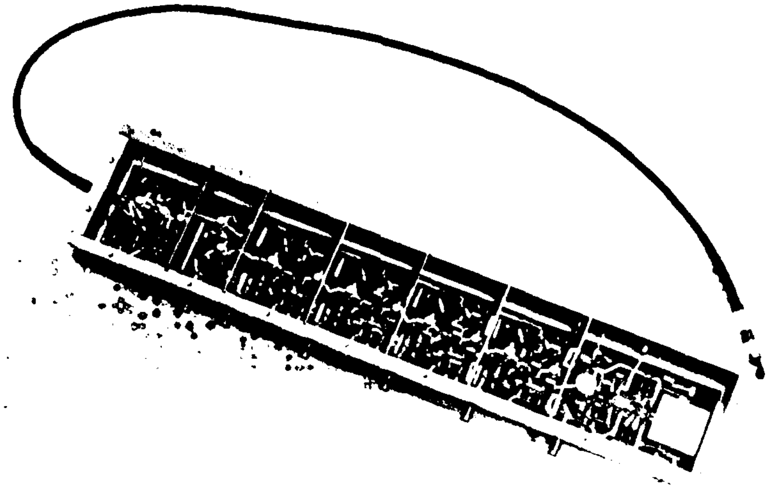


Figure 5-18. Receiver with Cover Removed

5.4.2 OPTICAL TRANSMITTER

The optical transmitter consisted of an off-the-shelf laser transmitter with additional buffer circuitry on a single PC card. The transmitter with the buffer cover removed is shown in Figure 5-19.

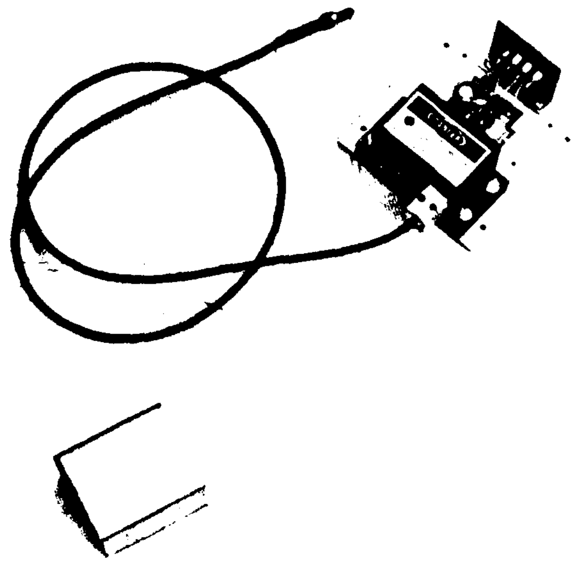


Figure 5-19. Transmitter with Buffer Cover Removed

The mechanical design simply provided the high-speed logic function and shielded the circuitry. No attempt was made to redesign the purchased analog laser transmitter; only adjustments were made. Laser noise levels were higher than desired but were adequate for a three- or four-terminal system. Designs optimized for digital applications are possible and are being pursued by vendors.

Figure 5-20 shows the complete transmitter and receiver units utilized in each terminal of the bus. The receiver and transmitter optical connections are made to the bus through the fiber-optic pigtails. The pigtails were actually connected to bulkhead connectors, one of which is shown in Figure 5-20, at a common bulkhead at the rear of the bus unit. Fiber-optic cable was then added to these connections to reach the multiport optical coupler located externally.

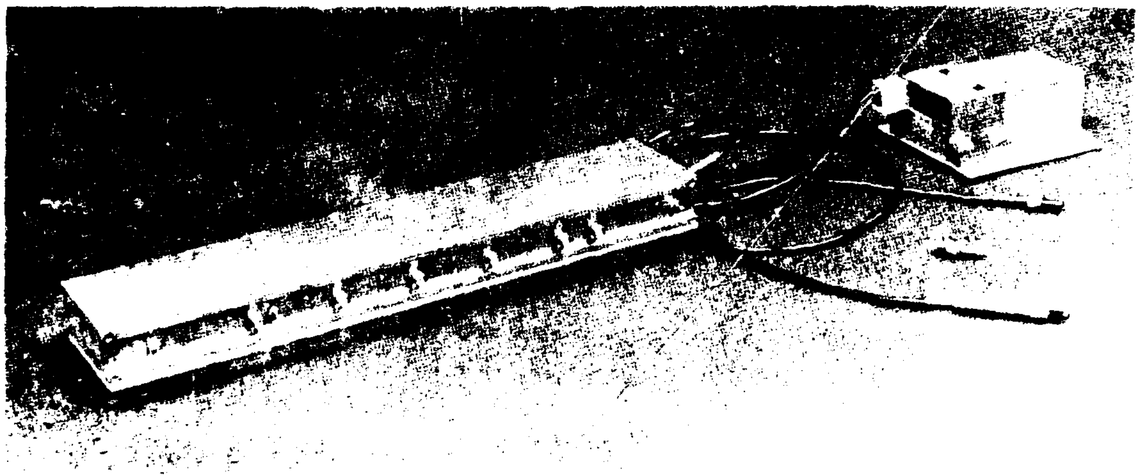


Figure 5-20. Complete Transmitter and Receiver Units

5.4.3 OPTICAL COUPLER

The 16-port transmissive optical coupler, Figure 5-21, was purchased from Spectronics Incorporated. It has 16 input terminals and 16 output terminals, measures roughly 5 x 5 x 1.5 inches, and has an aluminum housing. This small, lightweight unit represents an advance in packaging since previous programs. The coupler was intentionally located remote from the bus terminal units for the demonstration system, thus simulating installation in an actual system.

Figure 5-21. Optical Coupler

5.4.4 FIBER-OPTIC BUS COMPONENTS

All the components required for a terminal, its fiber interconnection, and the coupler are shown in Figure 5-22. The coupler and cable represent highly developed packaging. The receiver and transmitters utilize discrete components and do not represent advanced packaging. With advanced packaging and technology, the transmitter and receiver could likely be 10 times smaller.

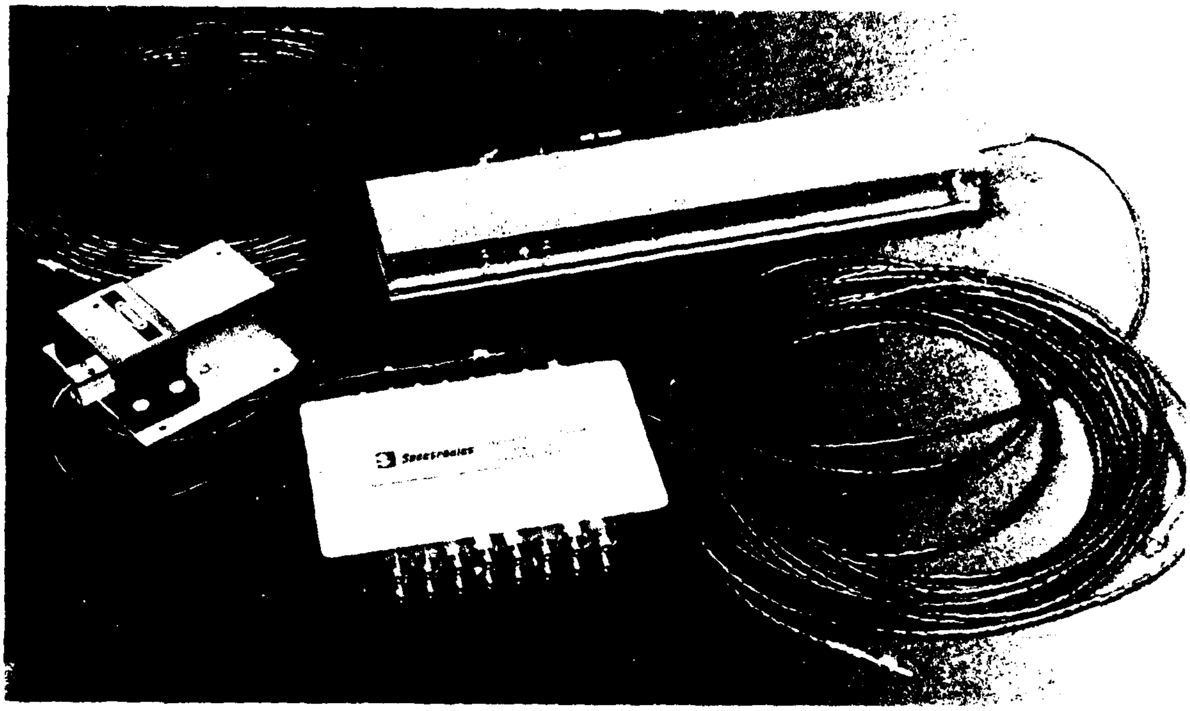


Figure 5-22. Set of Fiber-Optic Bus Components

5.5 FINAL SYSTEM SOFTWARE DESCRIPTION

The 50-Mb fiber-optic data bus demonstration system contains two software systems. One resides in the message operating system (MOS) and controls the three bus operating system (BOS) units. This controller-type program is written in Z80 assembler code.

The second software system is a collection of APL programs in two workspaces in the 5110. These programs perform the data generation, collection, and reduction required for the bit error rate/ incomplete message test and the bus access time/information rate test. Those tests were conducted as part of the final system test.

Three separate test programs were run at the system level. The first was named Terminal (Figure 5-23). This program allows only one message unit (terminal) to transmit data to itself on the fiber-optic bus. Terminal verified that each of the three message units could communicate properly on the fiber-optic bus.

The Bus program (Figure 5-24) is similar to the Terminal program but all three message units are attempting to access the bus.

Normal bus protocol is followed during this test. Any message unit may transmit to any other message unit. This program verifies proper bus operation.

The BAT program (Figure 5-25) conducts the bus access time and information rate tests. Only two message units are used: one is set up as a bus loader to simulate many terminals trying to access the bus, and the other simulates a single terminal but measures the terminal access time.

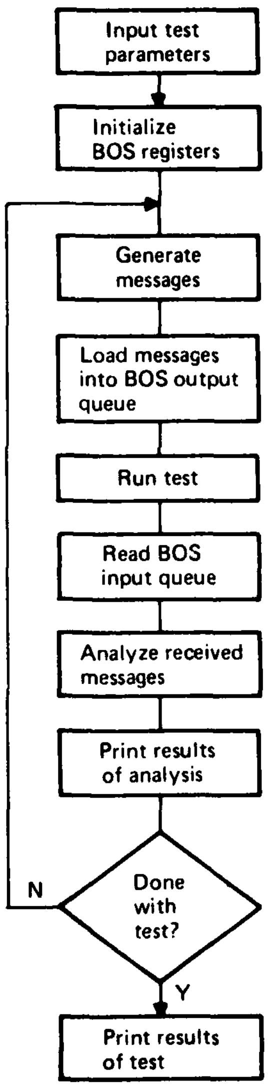


Figure 5-23. Terminal Flowchart

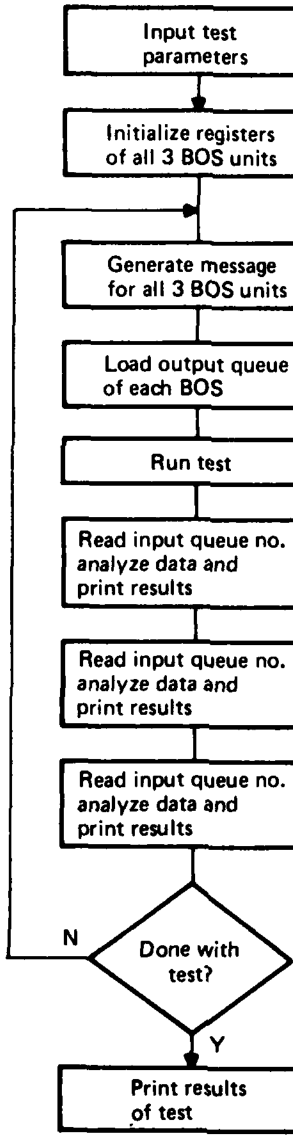


Figure 5-24. Bus Flowchart

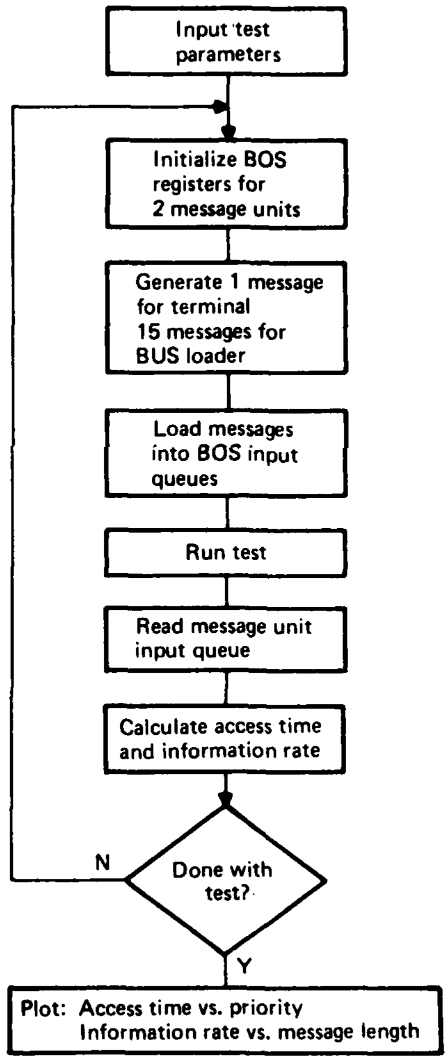


Figure 5-25. BAT Flowchart

5.5.1 TERMINAL PROGRAM

The Terminal Program routine (Figure 5-23) gathers and analyzes data for bit error rate and incomplete message rate measurements, and consists of a series of subroutines, each of which accomplishes some small part of the total task. Basically, Terminal asks for test parameters; sets up the message units; generates data; loads the queues with the data; issues the run command; reads the received data with status information from the BOS input queue, analyzes it for bit errors and incomplete messages, and then modifies the test parameters; and continues until the desired number of tests has been run. After the tests are complete, the results are printed and left in two variables for further analysis.

5.5.2 BUS PROGRAM

The Bus Program routine (Figure 5-24) conducts the same type tests as Terminal, but with the differences described next. All three message units are transferring data on the bus. Data can be transferred from any message unit to any other message unit, and, depending on the intermessage delay specified by the operator, messages may be interleaved on the bus.

5.5.3 BAT PROGRAM

The program to calculate and plot the bus access time measurements is called BAT, which sets up one message unit as a bus loader and one as a transmitting terminal (see Figure 5-25).

The transmitting terminal is assigned one message of a fixed priority to transmit continuously with a fixed intermessage delay. The bus loader has one message of every other priority and also has a fixed intermessage delay. When the transmitting terminal sends its message on the bus, no data goes into its input queue. Instead, the terminal loads the first word of the message into its queue, followed by the value of its internal access time counters.

When the input queue fills up, the 5110 reads this information and computes an average bus access time for this priority and bus loading conditions. Under program control, the transmitting terminal's priority is varied from minimum to maximum for each message length specified, and the results are stored for later use by the plot routines.

5.5.4 MOS SOFTWARE

An onboard EPROM contains the software that drives the MOS, a Z80 microprocessor-based message formatter/translater/controller. Commands from the 5110 are received at the MOS RS-232 port. The MOS decodes each command received; then jumps to a routine designed to handle the specified command. When a command is complete, control

is returned to the command decode section, and the MOS waits for the next command.

Under command from the 5110, the MOS can perform the following functions: (1) Load BOS control registers, including the test register, operation register, or intermessage-delay register; (2) load the BOS output queue with messages to be transmitted on the fiber-optic bus; (3) read BOS input queue, which will contain either messages received from the fiber-optic bus or access time measurements; (4) issue a Start command to the BOS units; and (5) read BOS Test Complete status and send it back to the 5110. See Figure 5-26.

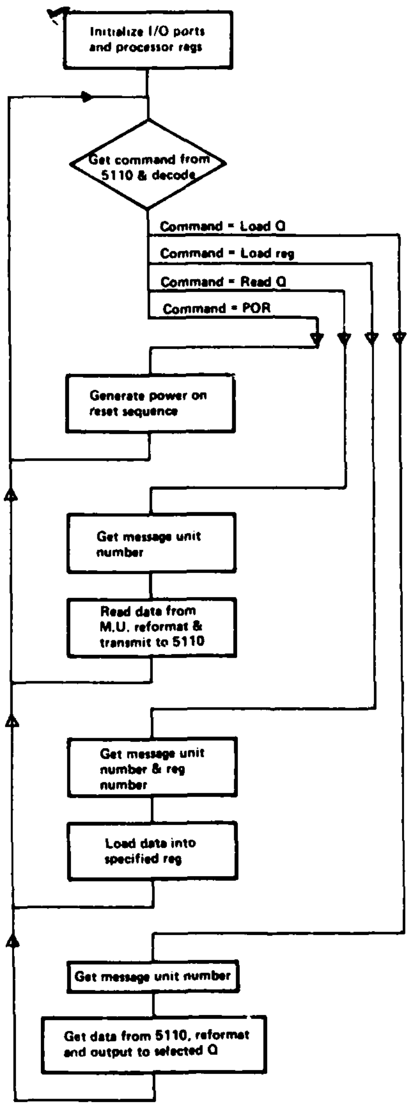


Figure 5-26. Z80 Software

5.6 TASK IIa TEST RESULTS

Contract data item C001, Fiber-Optic Final Demonstration System Development Plan, described the test plan to which the future aircraft fiber-optic interconnect system would be demonstrated and evaluated. These tests and demonstrations were concluded on 11 December 1979 at IBM Owego, NY. Three levels of tests were performed: component, subsystem, and final system. In some cases, test methods other than originally intended were utilized for better results, and in other cases, tests were not performed because of inaccessibility of test points in purchased items. In each case, deviations from the original plan are noted and explained.

5.6.1 COMPONENTS

5.6.1.1 Sources

Both LED and laser sources were purchased and measured. The laser source was part of a total transmitter unit supplied by General Optronics.

Table 5-2 gives the output power measurements of the sources and also the rise and fall times of the light output. Power measurements on the laser transmitter were taken at a transmitter supply voltage of -7 V and with no signal input. Power measurements on the LEDs were taken with a DC current of 100 mA. Laser transmitter rise and fall times were taken with a signal input of 0.8 V peak to peak, with high-speed emitter-coupled logic drivers.

The laser output rise and fall times were calculated from the formula

$$t_r = \sqrt{t_o^2 - t_i^2}$$

where t_o is the output rise or fall time, and t_i is the input rise or fall time

LED rise and fall times were measured using a high-speed (<2ns) pulse generator delivering a square wave current of 0 to 200 mA.

5.6.1.2 Detectors

Only one detector was ordered and measured. Its measured characteristics appear in Table 5-3.

The low responsivity resulted from a compromise to achieve the low depletion voltage.

Table 5-2. Output Power Measurements and Rise and Fall Times of Source

Light Output		
Source & Sample No.	Power (μW)	t_r/t_f (ns)
• Laser* (GOLT-3) outputs @ -7 V into transmitter package		
1	2,600	<1
2	2,460	<1
3	3,040	<1
4	2,300	<1
• Spectronics LED @ 100 mA		
1	820	12
2	1,080	12
3	660	12
• TI LED @ 100 mA		
1	78	5
2	78	5
3	68	5
4	68	5

*Readjusted all outputs to 2800 μW on 12/1/79

Table 5-3. Photodiode (Spectronics) Responsivity

Sample No.	Response (A/W)
1	0.26
2**	0.22
3	0.23
4	0.23
5	0.22
6	0.24
7	0.26

**Failed and replaced

5.6.1.3 Cables

Because of previous termination problems experienced by IBM and others, the originally recommended plastic-clad fibers were not ordered. Instead, a glass-clad silica fiber having an attenuation specification of 50 dB/km was procured in 10-m lengths with connectors.

Of 31 cables ordered and received, four were bad and seven failed later. Figure 5-27 shows the distribution of the attenuations of the original 27 good 10-m cables plus one connector.

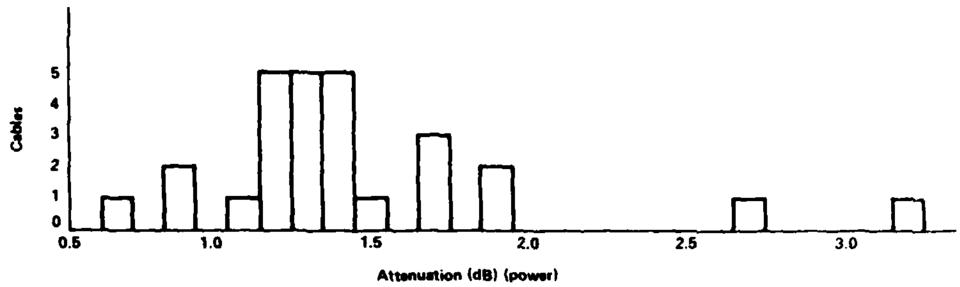


Figure 5-27. Distribution of Attenuation

Neglecting extremes, the average attenuation for a 10-m cable plus one connector is 1.3 dB. Because only 10-m cables were available, the separation of cable and connector attenuation was not possible. It is estimated that 1.0 dB of the attenuation is due to connectors and 0.3 dB to the cable. This translates to average cable attenuation of 30 dB/km.

5.6.1.4 Connectors

Amphenol Series 906 connectors were purchased and provided to the source, detector, cable, and coupler vendors to install on the pigtails and cable. In addition, ten compatible feed-through connectors were purchased by IBM.

From 5.6.1.3, it is estimated that the average feedthrough connector loss is 1.0 dB.

5.6.1.5 Multiport Couplers

Two types of 16-port passive transmissive couplers were received from Spectronics.

Figure 5-28 shows the distribution of port-to-port loss through the coupler with connectors, and Figure 5-29 is for the coupler with pigtails.

Only a representative sample of port-to-port measurements was made on the pigtailed coupler because many of the port-to-port paths had failed before delivery. Spectronics will replace this coupler at no charge.

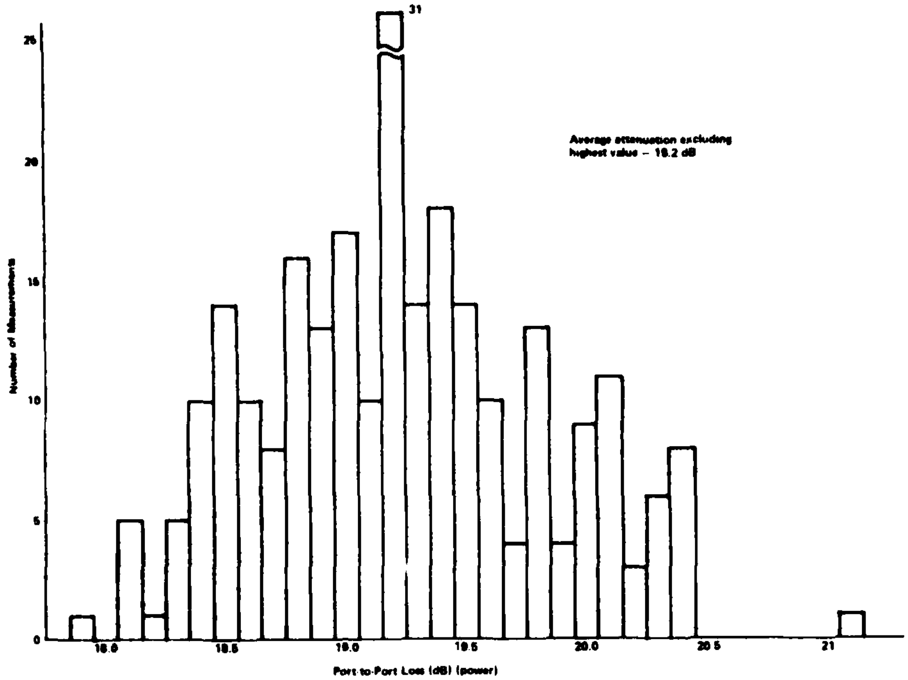


Figure 5-28. Port-to-Port Loss for Coupler With Connectors

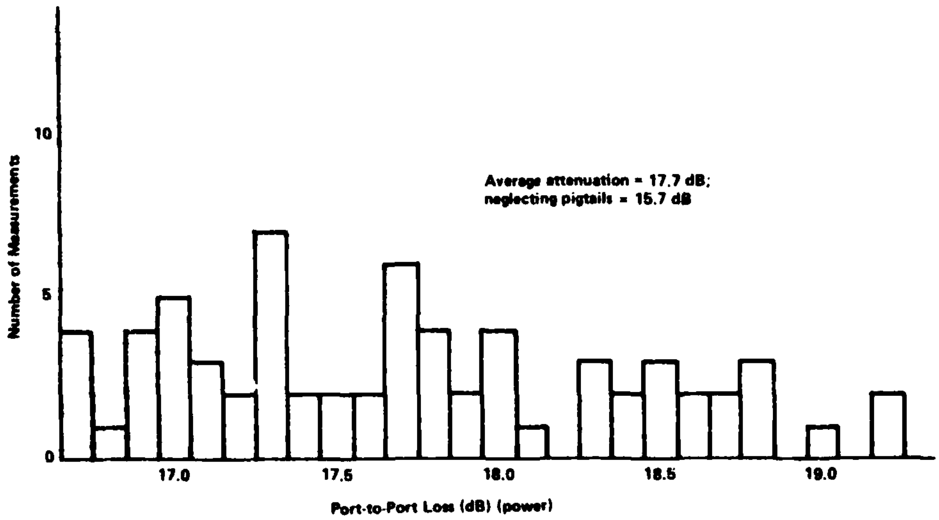


Figure 5-29. Port-to-Port Loss for Coupler with Pigtails

5.6.2 SUBSYSTEMS

5.6.2.1 Transmitter

Two types of transmitters were evaluated: a purchased laser transmitter, and an IBM-designed and fabricated LED transmitter.

5.6.2.1.1 Laser Transmitter

The General Optronics GOLT-3 laser transmitter was modified by installing a pigtail of the Gallite 3000 fiber cable with the Amphenol connector and by optimizing the internal circuitry for ECL signal level inputs of -0.8 to -1.6 V. A calibrated detector (integrating cube) was used to measure the peak light output of the transmitter at the end of the pigtail when the input signal was a low-frequency (5-kHz) square wave. The transmitter was not an optimum design for digital transmission because it was designed for amplitude- or frequency-modulated RF carrier applications.

The input signal to the transmitter is internally AC coupled by an RC network with a 5-ms time constant. When powered on in the absence of a signal, the laser is biased at a light power output level of 2,800 μW. An optical feedback with a time constant greater than 1 second attempts to maintain the average light output at this level in the presence of input signals.

At a 50-Mb Manchester-encoded data rate, messages of 2,000 words or less cause the laser output to vary between 2,800 μW and a level 5,000 μW higher (7,800), because the input time constant is too long to allow any seeking of average levels. For long messages (steady-state input), the signal will reach an average level and will swing between approximately 300 and 5,300 μW output. The actual transfer characteristic of the laser transmitter was 5.6 mW/V.

5.6.2.1.2 LED Transmitter

Because the system demonstration did not use LED transmitters, only a sampling of measurements was made. With the transmitter as designed and as described in 5.2.2.2, the output power was approximately 1 mW, with 10- to 12-ns rise and fall times.

5.6.2.2 Receiver

Several measurements were made on the receivers. Because it is basically an analog amplifier up to the comparators, which set and reset a latch to reconstruct the transmitted waveform, all the pertinent data was measured at the input to the comparator. It is at this point that signal-to-noise ratios, receiver sensitivity, signal levels, etc., are determined. Table 5-4 lists receiver measurements.

Table 5-4. Receiver Measurements

• Receiver sensitivity	-39 dBm
Noise (at comparator) (rms volts)	$\sqrt{(830 \times 10^{-6}) + \left(n \times 10^{-(\frac{A}{5} - 0.34)}\right)}$
A = system attenuation n = transmitters on line	
Signal (at comparator)	
At A = 18.9 dB	700 mV peak
At A = 27.9 dB	480 mV peak
Comparator threshold (bench)	250 mV
Actual noise (bench)	28.8 mV rms

Because of the laser transmitter characteristic (biased on with no signal), a noise contribution is present at the receiver which is a function of the attenuation on the bus and also the number of transmitters. The noise formula in Table 5-4 takes account of this but simplifies it by assuming the attenuation is equal for all transmitters. The comparator threshold and actual noise values are qualified by the term (bench). When the receiver was installed in the system, the threshold was raised because of the presence of three transmitters instead of one.

5.6.3 LINK TESTS

When the subsystems were operated as a link simulating a terminal-to-terminal transmission on a 16-port data bus, bit error rates (BER) were measured using the configuration of Figure 5-30. BER results for two levels of comparator thresholds are shown in Table 5-5. The 300-mV level was that used in system operation and tests.

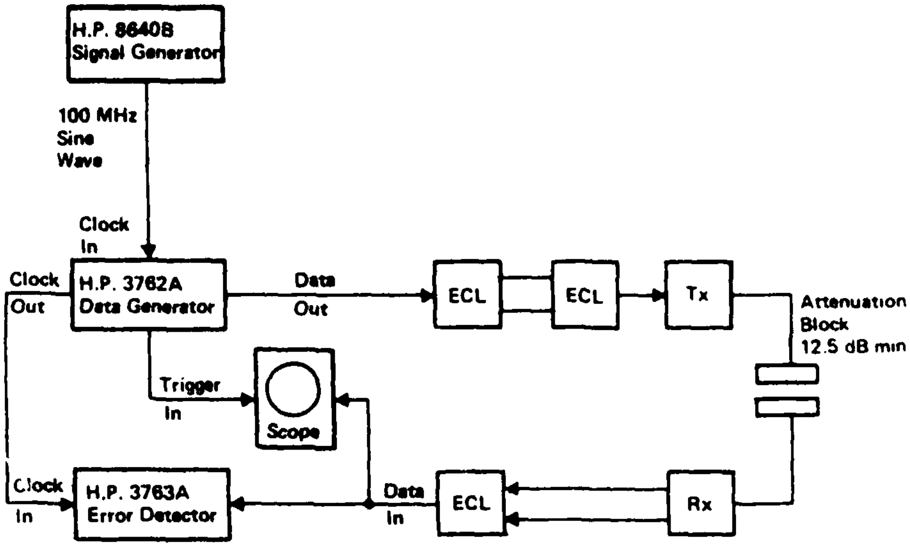


Figure 5-30. Subsystems Linked to Simulate Terminal-to-Terminal Transmission

Table 5-5. Link Test (receiver noise 28.8 mV rms)

Comparator Threshold (mV)	Attenuation (dB)	BER
300	29.1	1×10^{-4}
300	28.5	1×10^{-7}
300	27.9	$<1 \times 10^{-11}$
300	22.5	$<1 \times 10^{-12}$
250	29.1	2×10^{-7}
250	28.5	1×10^{-10}
250	27.9	$<1 \times 10^{-12}$

5.6.4 SYSTEM LEVEL

The system-level tests were divided into two sections. The first section was a special bus test used to measure the bit error rate and the incomplete message rate of a terminal. The second section was the operational bus test, used to measure the access time of a terminal and the protocol efficiency for the system.

The results of the tests are considered to be accurate and sound, even though there may have been some slight deviation from the original demonstration and System Development Test Plan.

5.6.4.1 Conclusions

The system-level testing performed on the high-speed fiber-optic data bus demonstration system supports three main conclusions:

- First, this testing showed that using our present technology, it is physically possible to construct and demonstrate a fiber-optic data bus with the following: asynchronous, free access, distributed control protocol; 50-MHz serial Manchester bit rate; 16 terminal ports; and a maximum terminal separation of 100 m.
- Second, our study has proved that this type of fiber-optic data bus is reliable with demonstrated bit error rates and incomplete word rates better than that required by the MIL-STD-1553B specification. The incomplete word rate and bit error rate were measured to be on the order of 1.4×10^{-8} and 6.4×10^{-10}, respectively.
- Third, this type of protocol's efficiency was examined with respect to bus access times and information rate. The total data bus information rate exceeded the system command and control bus requirements outlined in the initial Fiber-Optic Avionics Interconnect System Report (79-A77-001). This information rate exceeded 15 Mb/s for message lengths of 20 words and exceeded 30 Mb/s for message lengths of 100 words.

5.6.4.2 Special Bus Testing

As stated in the development and test plan, the purpose of the special bus test was to measure the bit error rate and the incomplete word rate of the system in the presence of noise.

5.6.4.2.1 Operational Test

The operational Bus program, which resides in the IBM 5110 (Figure 5-31) desk top computer, first initializes each of the three message units (MU) with a priority number and other operating modes. It then generates and distributes pseudo random data messages to each of these MUs. Next, the message operating system (MOS) issues a simultaneous run command, allowing the three MUs to function as independent terminals. Simultaneously, the MUs are competing for the data bus in order to pass their data messages to other terminals. After the three MUs have collected all of their messages, the MOS and 5110 interrupt the bus activity by reading and analyzing each of the three MUs. This is performed to determine whether that data resided in the proper MU and that there were no bit errors in the data transferred.

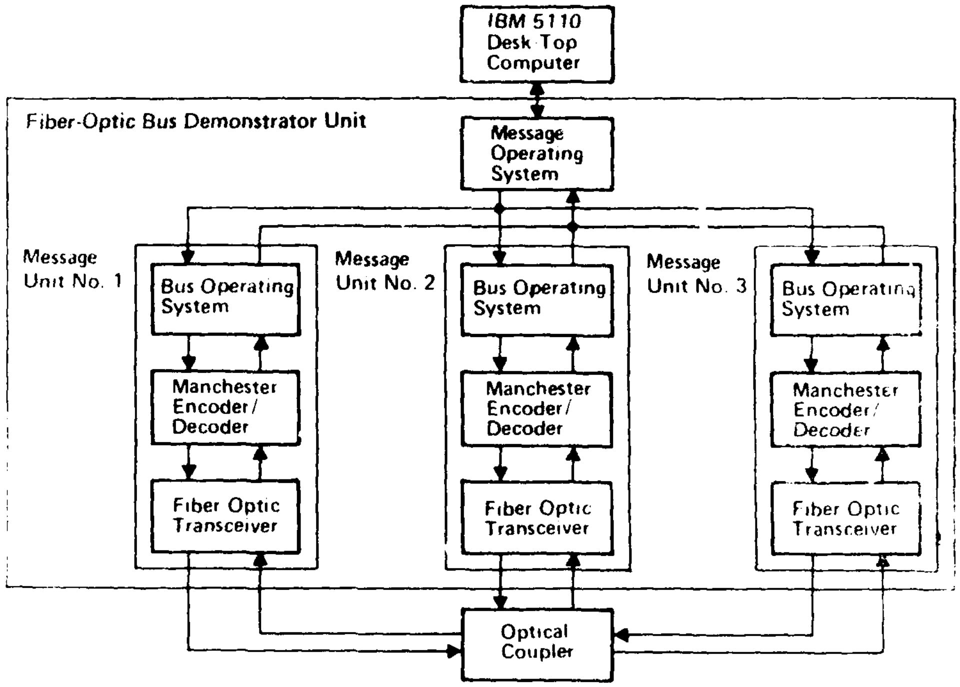


Figure 5-31. High-Speed Fiber-Optic Data Bus Demonstration System

Table 5-6 indicates the conditions and results of the Bus operating program. This program was very important in the verification of total simultaneous operation of the Distributed Control Protocol Data Bus. It is evident, though, that this program was lacking in the amount of exchanged data and necessary time to generate and analyze this data.

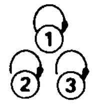
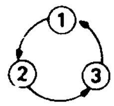
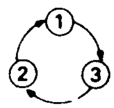
5.6.4.2.2 Bit Error Rate and Incomplete Word Rate Test

The test program 'Terminal' was used with internally built test equipment called the system error rate counter to measure the bit error rate and incomplete word rate of the data bus system.

The Terminal test program is a subset of the operational Bus test program because it performs identical functions. However, the program is limited to only one communication path at a time; for example, a message transfer between MU 1 and MU 3. This program has been further modified to instruct the transmitting MU to continually send its output queue's contents via a wrap feature of the bus operating system.

The system error rate counter (Figure 5-32) is simply three decimal counters which can be attached to count any TTL signal and which can be multiplexed to a 12-digit decimal display.

Table 5-6. Results of Bus Testing

	Message Unit	Conditions: Message Length Minimum/Maximum/Increment in Words	Words Transferred per Test	Time per Test (h)
Case 1		5/50/5 50/250/25	20,000	1
Case 2		2/250/2	200,000	6
Case 3		No detected errors		

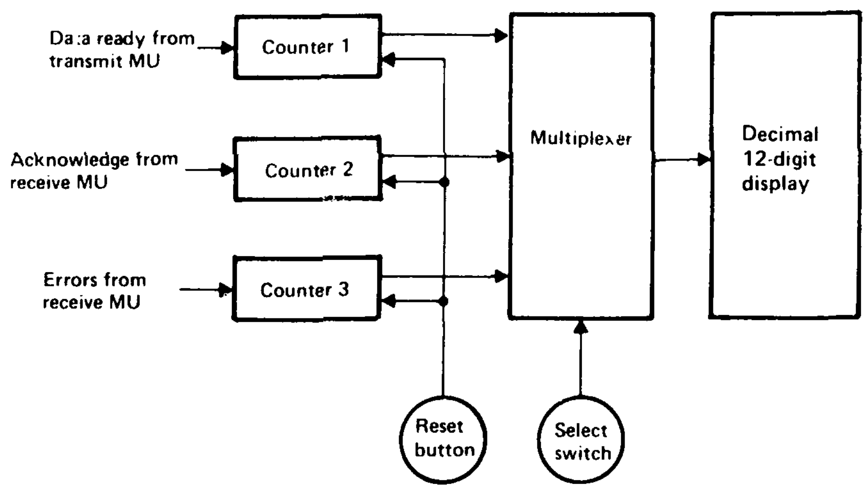


Figure 5-32. System Error Rate Counter

With the Terminal test program generating data and the system error rate counter attached to appropriate logic circuits (indicating system activity and system errors), bit error rates and incomplete word rates were measured for each of the nine possible message unit communication paths. Over one billion data words were transferred for each test.

The results of this activity are presented in Table 5-7, with the final average of $\sim 6.4 \times 10^{-10}$ for the system bit error rate (BER) and $\sim 1.4 \times 10^{-8}$ for the system incomplete word rate (IWR).

The case with message unit 2 transmitting to MU 1 is especially interesting because its BER and IWR are considerably worse than for any other case. In other cases, these same MUs performed well. This indicates that the total noise and/or dB loss in this particular communication path are greater than for the other paths.

Table 5-7. Fiber-Optic Data Bus Error Detection

Message Unit XMIT	RCVR	Bit Error Rate ($\times 10^{-10}$)	Incomplete Word Rate ($\times 10^{-9}$)
1	1	6.25	1.80
1	2	1.75	5.50
1	3	4.75	10.50
2	1	22.50	46.50
2	2	0.00	0.00
2	3	1.75	3.50
3	1	6.25	1.25
3	2	2.00	4.50
3	3	12.50	50.00
Average		6.4×10^{-10}	1.4×10^{-8}

$$\text{Bit error rate} = \frac{\text{No. incorrect bits received}}{20 \times \text{No. words transmitted}}$$

$$\text{Incomplete word rate} = \frac{\text{no. words with incorrect bits received and no. words missed}}{\text{no. words transmitted}}$$

5.6.4.3 Operational Bus Testing

The operational bus testing measured the overall efficiency of transmitting information between terminals using a distributed control protocol. Bus access times were also measured for different terminal priority numbers and for different amounts of system bus loading.

The Bus Access Time (BAT) software program, in conjunction with a Plot routine, was used to collect and plot the bus access times and information rate results.

The BAT program initializes the message units and generates data messages in the same manner as the Bus and Terminal programs but differs from the latter two programs in two basic areas. First, one of the message units is set up with messages of ascending priorities. Second, the terminal analyzed is preprogrammed to fill its input queue with bus access times. These are read in from special counters built into the bus operating system rather than from data messages transferred over the bus.

The FO data bus demonstration unit and its operational programs are very versatile. They can allow for an infinite combination of data runs. Four plots summarize the bus access time and information rate results. The rate data for these plots is in the appendix, and the plots are described next.

5.6.4.3.1 Access Time as a Function of Test Number

Figure 5-33 shows the best possible access time a terminal can have as a function of the terminal's priority number with no bus loading. Priority 0 is the highest; priority 15 is the lowest. All these points lie on a straight line, with a Y-axis shift of approximately 500 ns. This plot could be summarized with the equation

$$\text{Terminal access time} \approx (\text{priority no.}) \times (\text{access window} \approx 1.25\ \mu s) + (\text{bus response} \approx 0.5\ \mu s)$$

For example, if a message unit had a priority number of 8 and there was no data bus contention, it would take 10.4 μs to access the bus.

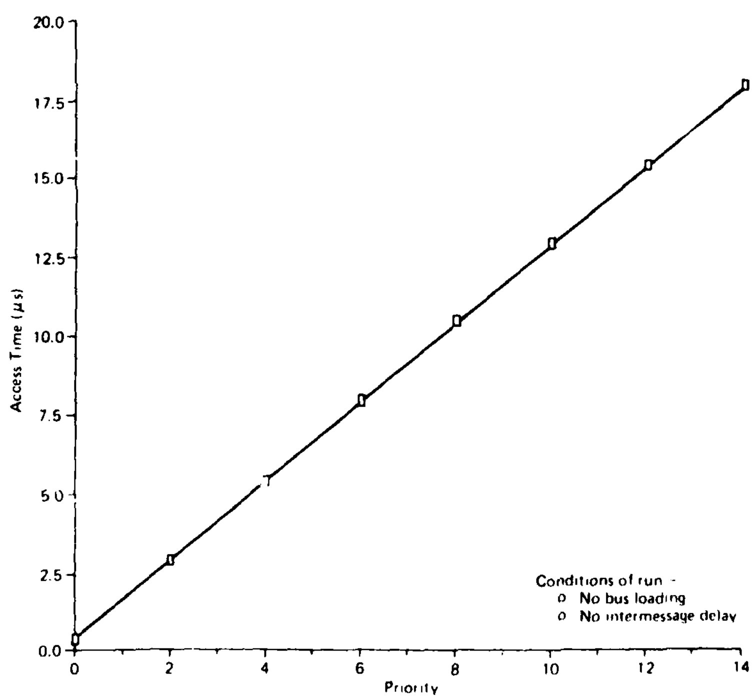


Figure 5-33. Access Time as a Function of Priority for no Bus Loading

It should be emphasized that the bus operating system (BOS) is designed for a maximum terminal separation of 100 m, which directly affects the access window time. If a system is desired with a smaller maximum terminal separation, the BOS could be modified to run with a smaller bus access window.

Figure 5-34 is a worst-case condition of terminal access time as a function of terminal priority number with 100% bus loading. For example, if all 16 terminals, each with a different priority number, were simultaneously trying to access the bus with a 20-word message, and a message unit had a priority number of 8, it would take 113.12 μs to access the bus. Note that if there is 100% bus loading with priority numbers 0-14, then a message unit with a priority number of 15 will never access the bus. Obviously, when a system is designed around this type of bus architecture, much consideration must be given to understand the bus loading of every priority of MU. Therefore, no one MU can be locked off the bus.

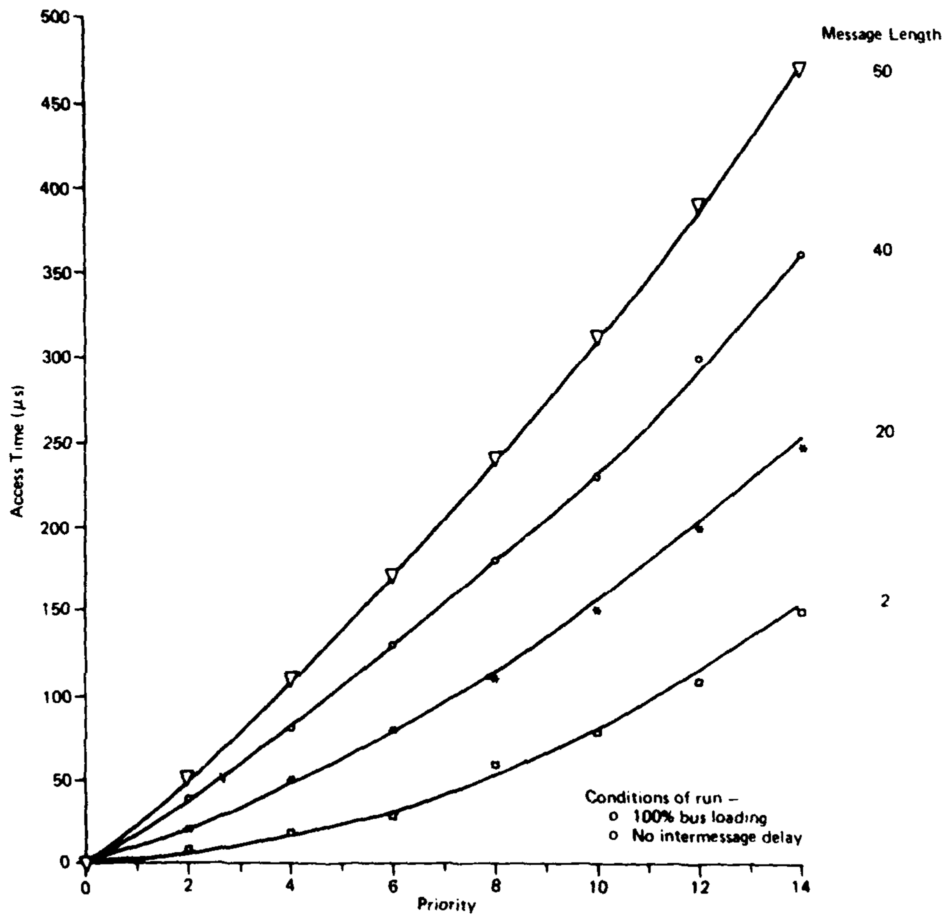


Figure 5-34. Access Time as a Function of Priority for 100% Bus Loading

5.6.4.3.2 Information Rate As A Function of Message Length

Figure 5-35 is a best-case condition of terminal information rate as a function of message length with no bus loading. The priority 0 curve depicts the best possible bus information rate if the system is set up for a maximum terminal separation of 100 m. The priority 8 curve was used to indicate the average bus information rate. The dotted line is a system prediction of the average bus information rate, which compared very closely with that actually measured. The difference between the two is that the predicted bus access time is closer to BAT = (BW × P) + bus response time

where

Bus response time ≈ 500 ns

The results of these plots can be summarized by the following equation:

$$\text{Information rate} = \frac{(16 \text{ information bits}) \times ML}{BAT + (20 \text{ total bits}) \times T \times ML}$$

where ML = message length

BAT = either measured or predicted by (BW) × P + bus response time

BW = bus window time = 1.25 μs

P = priority number

T = bit transfer time = 0.02×10^{-6}

For example, if a MU had a priority of 8 and was sending 100-word messages with no bus loading, that device would have an information rate of 31.75 Mb/s.

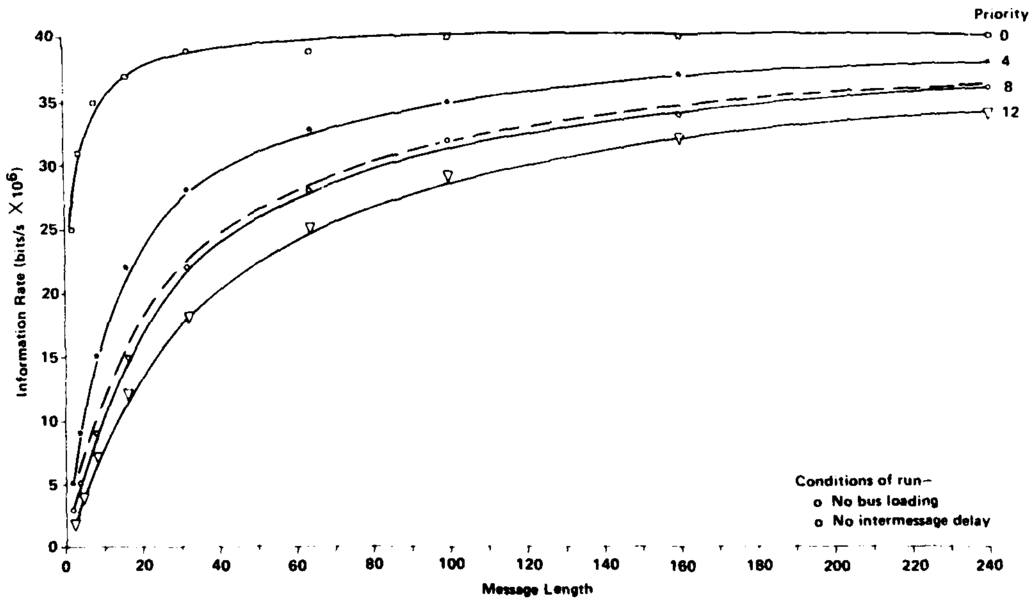


Figure 5-35. Information Rate as a Function of Message Length for No Bus Loading (Best Case)

Figure 5-36 is a worst-case condition of terminal information rate as a function of message length with 100% bus loading. For example, if a MU with priority of 8 wishes to transmit a 100-word message, but has to wait for message units with priorities 0-7 each with a 20-word message, the terminal information rate is 10.44 Mb/s.

Naturally, the terminal information rate is lowered with bus loading, and this is just one specific condition. Note that the degree of drop in information rate is smaller for higher priority MUs, and in this case, a MU with priority of 0 is not affected by bus loading.

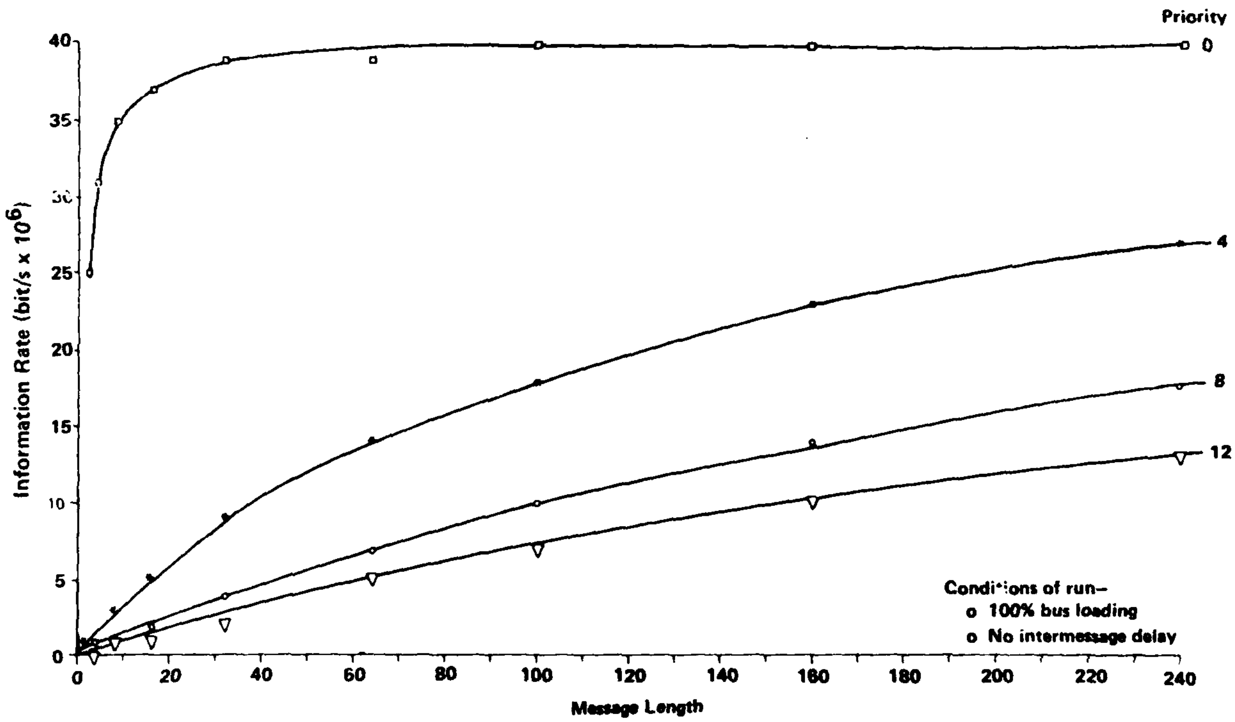


Figure 5-36. Information Rate as a Function of Message Length for 100% Bus Loading (Worst Case)

Appendix A
BIT ERROR RATE COMPARISON - PPM VERSUS FULL-WIDTH MANCHESTER FOR FIBER-OPTIC 1553B COMPATIBLE DATA BUS

Section 3.4.4 developed a comparison between pulse position modulated (PPM) optical Manchester signals and 1553B electrical full-width Manchester in terms of achievable bit error rates (BERs). It was shown that the logic-level Manchester signal reconstructed from the PPM optical signal was not affected by rise time or jitter problems in determining the system BER capability.

Also, in the decoding circuitry, the sampling time for each half bit would always be within the limits of the reconstructed waveform. Finally, it was shown that if full-width Manchester was used, the amplitude of the bus signal must be increased to account for the effects of rise time (frequency response) and jitter.

Figure A-1 is a computer-generated waveform for a 1-Mb, full-width Manchester-encoded signal (sync plus first two bits) after passing through a low-pass filter with two poles at 2 MHz. The sampling time for each data half bit is synchronized by the time at which the center of the sync pulse crosses the detection threshold.

It is apparent that any noise on the signal transition at this time will make this "zero crossing" ambiguous by some amount, with the probability for that amount determined by the signal-to-noise ratio (with respect to the threshold) at the corresponding signal amplitude on the waveform transition. In Figure A-2, this kind of ambiguity is shown as a cross-hatched area between the +A and -A levels for both the zero crossing and for the pulse width of the first half of the first data bit.

The probability for this particular ambiguity would be determined by the S/N ratio at +A or -A. Because the edge of the sampling clock can fall anywhere (relative to the center of the first bit half) within the amount of zero crossing ambiguity, the pulse to be sampled must have beginning and ending times which include this ambiguity; that is, in Figure A-2, $Wz \leqslant Wp$.

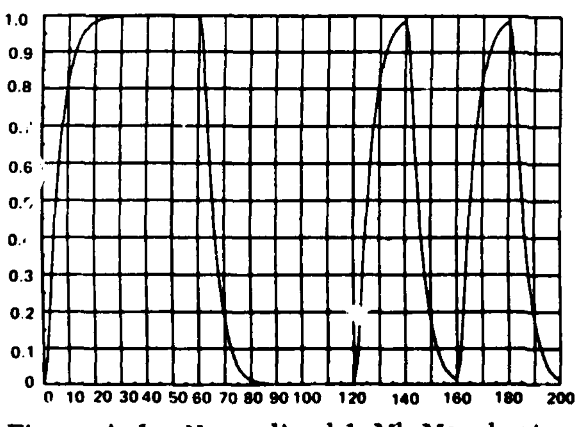


Figure A-1. Normalized 1-Mb Manchester (0-2 MHz) (Arbitrary Time Scale)

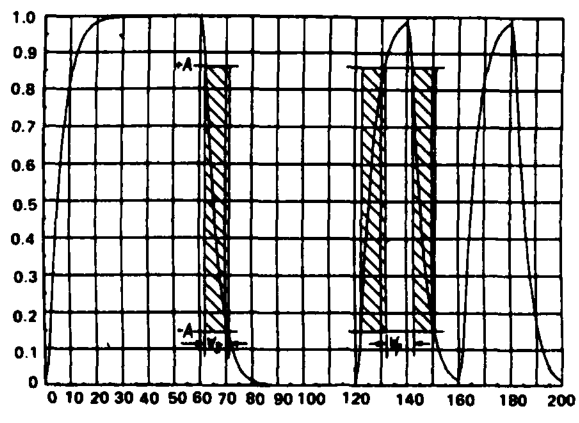


Figure A-2. Ambiguity for Zero Crossing and Pulse Width (0-2 MHz)

In a well-designed receiver, one would like equal probabilities that the reconstructed data pulse widths and zero crossing ambiguity are the same. Figure A-3 shows a graphical method for ensuring this, and it also allows the determination of the signal amplitude requirements for a given BER. In Figure A-3, if the sync transition were perfect (zero fall time and no jitter), it would be timed at 60 on the relative horizontal scale. The sampling time synchronized to this zero crossing would occur at 130 on the same scale for the first half of the first data bit. With a noninfinite bandwidth, the waveform transitions will not take zero time, and they will have jitter with respect to one another.

If one "translates" the limit curves of the mid-sync transition to the point where the center of the first half bit should be (see Figure A-3), one can determine at what normalized amplitude level the probability for the reconstructed pulse width and zero crossing ambiguity durations will coincide. This amplitude is where the translated limit transitions of the sync cross the rise and fall edges of the first half bit of the first (or any) data pulse. Figure A-4 is an expanded view of this first half bit for a 2-MHz upper cutoff frequency and a ±25-ns jitter. At the amplitude where either sync transition crosses the pulse such that the sample time will always be within the pulse period, the S/N ratio must be great enough to achieve the required system bit error rate.

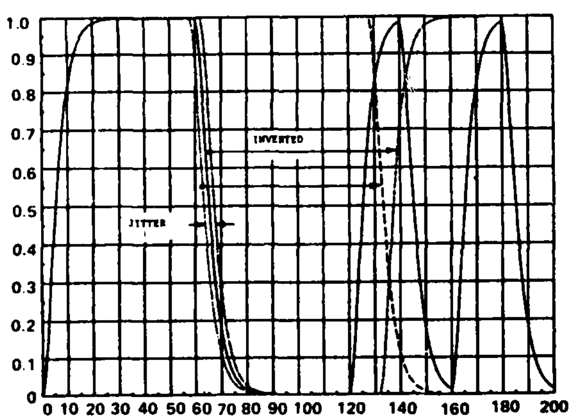


Figure A-3. Technique to Determine Amplitude Requirements For a Given Bit Error Rate

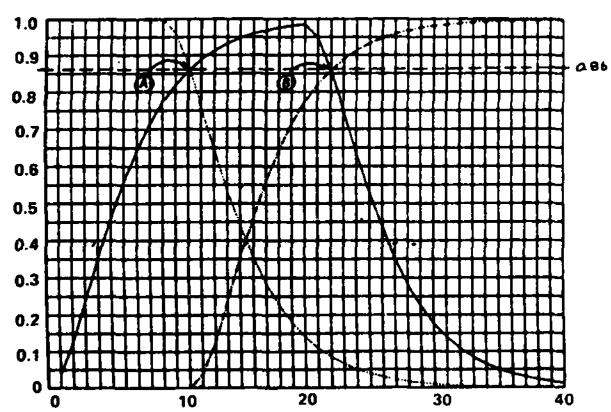


Figure A-4. Expanded First Half Bit (0-2 MHz, 25-ns Jitter)

Refer again to Figure A-4. The probability that the reconstructed pulse will be between the two transition crossings (A and B) is

$$P_p = 1 - \text{erfc}\{SNR\}*$$

where SNR is the signal-to-threshold-noise ratio at the crossing amplitude. The probability that the sync zero crossing will be between these limits is the same; that is,

$$P_z = 1 - \text{erfc}\{SNR\}$$

The probability that both events occur together is

$$P_p \cdot P_z = 1 - 2\,\text{erfc}\{SNR\} + (\text{erfc}\{SNR\})^2$$

In the range of values considered,

$$(\text{erfc}\{SNR\})^2 << 2\,\text{erfc}\{SNR\}$$

Therefore, the probability for a good sampling is

$$P_s = 1 - 2\,(\text{erfc}\{SNR\}$$

The probability of an error is

$$BER = 2\,\text{erfc}\{SNR\}$$

If one desires a BER of 1×10^{-9}, then,

$$2\,\text{erfc}\{SNR\} = 1 \times 10^{-9},$$

$$\text{erfc}\{SNR\} = 0.5 \times 10^{-9}$$

$$SNR = 6.12$$

In Figure A-4, 0.86 is the normalized amplitude at which the zero crossing ambiguity and the minimum pulse width have equal probabilities. The detection threshold in the receiver is at 0.5. The peak-to-peak normalized amplitude around threshold is $2 \times (0.86 - 0.5) = 0.72$.

The total pulse peak-to-peak value must be

$$2 \times 6.12 \times \frac{1}{0.72} \times A_n$$

for a BER of 1×10^{-9}, where A_n is the rms noise current (or voltage) amplitude, whatever the units may be.

For full-width Manchester, for the BER of 1×10^{-9}, with an upper frequency cutoff of 2 MHz and ±25 ns jitter, the peak-to-peak signal must then be 17 times the rms noise; that is,

$$S = 17 \times A_n$$

$$*\text{erfc}\, x = \frac{1}{\sqrt{2\pi}} \int_x^{\infty} \epsilon^{-\frac{t^2}{2}}\, dt$$ (complimentary error function)

Table A-1 shows the effect of upper frequency cutoff and jitter on this number. Figures A-5 through A-9 show the computer-generated curves for the other parameter combinations of Table A-1.

Table A-1. Peak-to-Peak Signal to Noise Ratios Required for BER = 1 x 10^{-9}

	Full-Width Receiver Cutoff Frequency (Hz)		
Jitter (ns)	1.5 x 10^6	2 x 10^6	3 x 10^6
±25	21.5	17	13.9
±50	24.0	18.8	14.6

To compare PPM with full-width Manchester, one must examine the receiver designs for each. If one assumes a transimpedance preamplifier with an FET front end, the largest value one can use for the feedback resistor depends on the shunt capacity of that resistor and the upper cutoff frequency required for that particular signal.

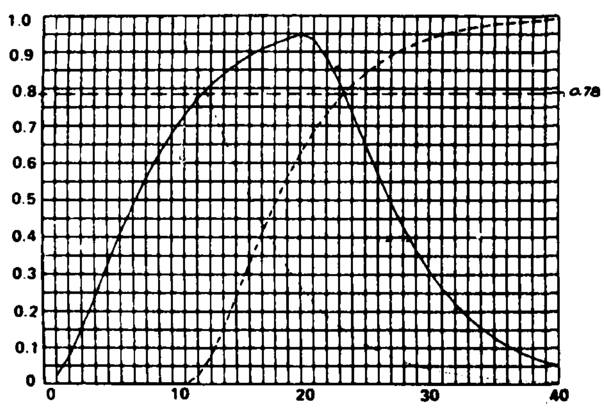


Figure A-5. Expanded First Half Bit (0-1.5 MHz, 25-ns Jitter)

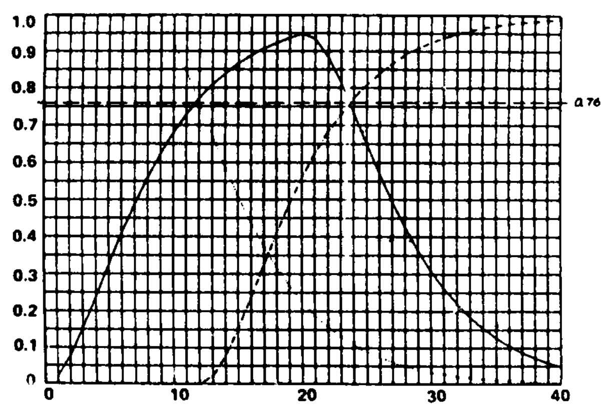


Figure A-6. Expanded First Half Bit (0-1.5 MHz, 50-ns Jitter)

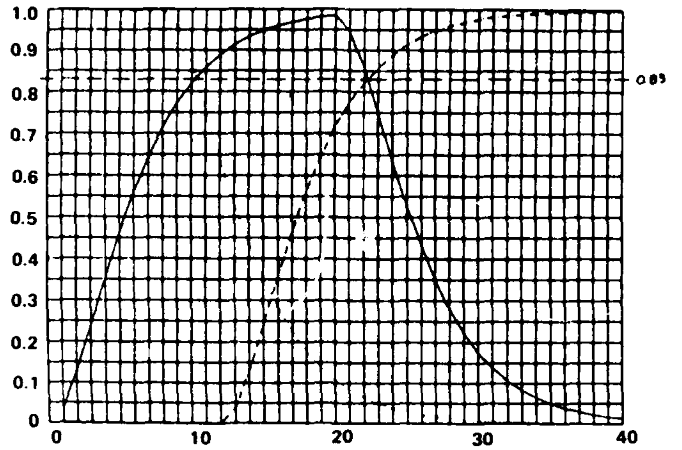


Figure A-7. Expanded First Half Bit
(0-2MHz, 50-ns Jitter)

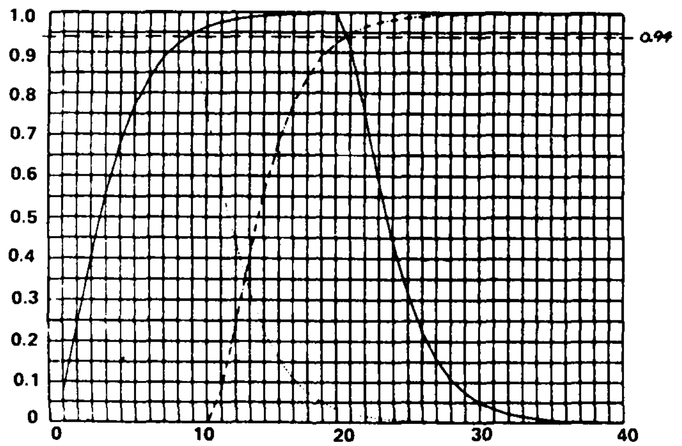


Figure A-8. Expanded First Half Bit
(0-3 MHz, 25-ns Jitter)

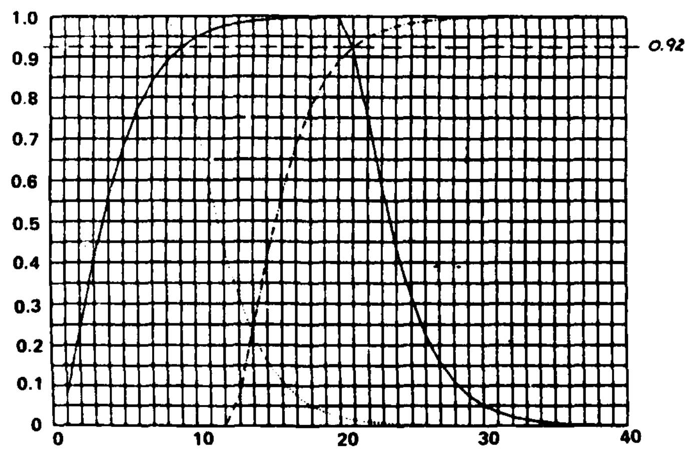


Figure A-9. Expanded First Half Bit
(0-3 MHz, 50-ns Jitter)

If we assume a receiver with a bandpass filter with one low-frequency pole at f_L and two high-frequency poles at f_H, then f_H can be considered the high-frequency cutoff (3 dB down) of the preamp, and the other poles at f_L and f_H are in the postamp. If α is the ratio of f_L to f_H, one can use the following equation to determine the equivalent input rms noise current for that amplifier at room temperature:

$$\frac{i_n}{\sqrt{H_z}} = 1.28 \times 10^{-10} \sqrt{\frac{1}{R_f} + \frac{C_T^2 \; B_n^2}{0.023 \; g_m} \; \frac{(1-\alpha^2)^4 \; (1+2\alpha)}{(1-\alpha)^4}} \; [1]$$

where C_T is the input shunt capacitance, R_f is the feedback resistor, g_m is the transconductance of the FET, and B_n is the noise bandwidth.

$$B_n = \frac{\pi f_H}{4} \left(\frac{1-\alpha}{1-\alpha^2} \right)^{2\,*}$$

The factor 0.023 is a characteristic of the FET configuration and the double upper pole. Table A-2 shows the calculations for rms noise for the configurations examined and assumes the $C_T = 5 \times 10^{-12}$ and $g_m = 10^{-3}$.

Table A-2. Preamplifier Noise

Parameter	PPM (α=0.125)	Full Width ($\alpha \simeq 0$)		
Upper cutoff frequency (Hz)	8×10^6	1.5×10^6	2×10^6	3×10^6
Shunt C across R_f (F)	0.3×10^{-12}	0.3×10^{-12}	0.3×10^{-12}	0.3×10^{-12}
R_f (kΩ)	66	350	270	180
$i_n/\sqrt{Hz}$	*1.06×10^{-12}	0.268×10^{-12}	0.322×10^{-12}	0.435×10^{-12}
$i_{n(rms)}$ (A)	2.36×10^{-9}	0.292×10^{-9}	0.403×10^{-9}	0.670×10^{-9}

*Although the upper cutoff frequency is 8 MHz, the lower frequency cutoff is 1 MHz and the resulting B_n is therefore 4.96×10^6 Hz.

It has previously been shown (Task I) that the optical peak pulse power for PPM can be 10 times that for full-width Manchester. We have also calculated the required peak-to-peak S/N ratios required for a BER of 10^{-9} for full-width Manchester (Table A-1). Table A-3

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[1] This and other equations are developed in Appendix B.

shows the actual signal input current amplitudes required for the full-width Manchester (combination of Table A-1 and Table A-2); that is, $i_s = SNR \times i_n$, for a BER of 1×10^{-9}.

Table A-3. Peak-to-Peak Signal Required (i_s) (A)

	Cutoff Frequency (Hz)		
Jitter (ns)	1.5×10^6	2×10^6	3×10^6
±25	6.3×10^{-9}	6.9×10^{-9}	9.3×10^{-9}
±50	7.0×10^{-9}	7.6×10^{-9}	9.8×10^{-9}

For a PPM peak-to-peak signal 10 times greater than the full-width Manchester, and combining the rms noise for PPM from Table A-2 and the results from Table A-3, Table A-4 gives the resulting peak-to-peak SNRs for PPM when compared with the full-width Manchester. Remember that the full-width numbers are required for a BER of 10^{-9}. Table A-5 converts the SNRs to BERs if the PPM threshold is assumed to also be at the half-amplitude level of the pulse.

Table A-4. Peak-to-Peak SNR

		Full-Width Cutoff Frequency (Hz)		
Jitter (ns)	Mode	1.5×10^6	2×10^6	3×10^6
±25	FW	21.5	17	13.9
	PPM	26.7	29.2	39.4
±50	FW	24.0	18.8	14.6
	PPM	29.7	32.2	41.5

Table A-5. Bit Error Rates

		Full-Width Cutoff Frequency (Hz)		
Jitter (ns)	Mode	1.5×10^6	2×10^6	3×10^6
±25	FW	1×10^{-9}	1×10^{-9}	1×10^{-9}
	PPM	5.9×10^{-41}	1.4×10^{-48}	1.08×10^{-86}
±50	FW	1×10^{-9}	1×10^{-9}	1×10^{-9}
	PPM	3.5×10^{-50}	1.28×10^{-58}	6.1×10^{-96}

Table A-5 tends to exaggerate the practical difference between PPM and full-width Manchester because of the nature of the complementary error function. Looking at the differences from another view, note the greater attenuation allowed in the optical link when using PPM. From the complementary error function tables, a peak-to-threshold SNR of 6 is required for a BER of 1×10^{-9} when using PPM. A peak-to-peak of 12 results, and if the PPM peak-to-peak SNRs of Table A-4 are divided by 12 and converted to dB, Table A-6 evolves. (In this case, dB = $10 \log \frac{SNR}{12}$).

Table A-6. Additional Attenuation Allowable for PPM (dB)

	Full-Width Cutoff Frequency (Hz)		
Jitter (ns)	1.5×10^6	2×10^6	3×10^6
±25	3.5	3.9	5.2
±50	3.9	4.3	5.4

This analysis assumes a sampling time perfectly synchronized with the apparent zero crossing of mid sync. This rarely happens and is another source of error that could degrade the BER even further.

Some decoding schemes resynchronize the sampling time at each waveform transition. Although the analysis changes slightly, the results are identical when examining sequential bits whose transitions can be in error in the same way the sync zero crossing and a sampled bit are in this analysis.

A further advantage for the PPM encoding can be achieved by using a bipolar transistor front end transimpedance amplifier. This configuration for an upper frequency cutoff of 8 MHz is less noisy than the JFET front end. In Table A-2, the i_n (rms) for PPM would be 1.95×10^{-9} instead of 2.36×10^{-9}. This would add an advantage of 0.8 dB to each of the numbers in Table A-6.

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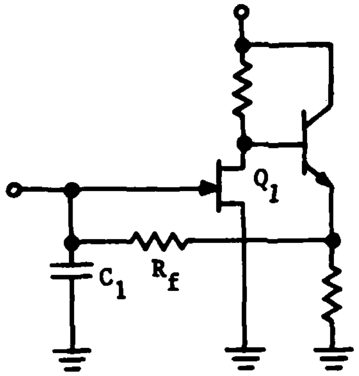
Appendix B
DEVELOPMENT OF EQUATIONS

B.1 DEVELOPMENT OF EQUATION FOR NOISE IN A JFET FRONT END TRANSIMPEDANCE AMPLIFIER

For the circuit,



Q_1 = 2N4417
$g_m = 10^{-3}$
C_1 = 5 pF (total input C)
C_{gs} = 0.8 pF (gate to source C)

the equivalent noise input current is obtained by integrating the noise spectral density, $\Phi(\omega)$, modified by the receiver filter transfer function, $G(j\omega)$.

$$\overline{i_n^{\,2}} = \frac{1}{2\pi} \int_0^{\infty} \Phi(\omega) / G(j\omega) /^2 d\omega$$

where $\overline{i_n^{\,2}}$ is the noise input average for a one-sided*noise bandwidth $B_n(\omega)$. For the JFET preamp,

$$^{**}\quad \Phi(\omega) = \frac{4kT}{R_f} + \frac{4kT\gamma\omega^2 C_1^{\,2}}{g_m}$$

where $\gamma = 2/3$ for a JFET [1]

---

*No negative frequencies

**For $\frac{C_{gs}}{C_1}$ = 0.16, drain thermal noise predominates (gate noise neglected)

[1] VanderZiel, Noise in Measurements, 1976 John Wiley and Sons, p 54.

If one considers a bandpass filter with a low-frequency pole (-3 dB) at f_L and two high-frequency poles at f_H,

$$|G(j\omega)|^2 = \frac{x^2}{(x^2+\alpha^2)(x^2+1)^2} \qquad \text{where } x = \frac{\omega}{\omega_H} \quad \text{and } \alpha = \frac{\omega_L}{\omega_H}$$

then,

$$\overline{i_n^2} = \frac{4kT}{2\pi}\left\{\frac{\omega_H}{R_f}\int_0^\infty \frac{x^2dx}{(x^2+\alpha^2)(x^2+1)^2} + \frac{\omega_H^3 C_1 \gamma}{g_m}\int_0^\infty \frac{x^4dx}{(x^2+\alpha^2)(x^2+1)^2}\right\}$$

Expanding in partial fractions,

$$\frac{1}{(x^2+\alpha^2)(x^2+1)^2} = \frac{1}{(1-\alpha^2)^2}\left[\frac{1}{x^2+\alpha^2} - \frac{1-\alpha^2}{(x^2+1)^2} - \frac{1}{x^2+1}\right]$$

then,

$$\omega_H\int_0^\infty \frac{x^2dx}{(x^2+\alpha^2)(x^2+1)^2} = \frac{\omega_H}{(1-\alpha^2)^2}\left\{\int_0^\infty \frac{x^2dx}{x^2+\alpha^2} - (1-\alpha^2)\int_0^\infty \frac{x^2dx}{(x^2+1)^2} - \int_0^\infty \frac{x^2dx}{x^2+1}\right\}$$

and,

$$\omega_H^3\int_0^\infty \frac{x^4dx}{(x^2+\alpha^2)(x^2+1)^2} = \frac{\omega_H^3}{(1-\alpha^2)^2}\left\{\int_0^\infty \frac{x^4dx}{x^2+\alpha^2} - (1-\alpha^2)\int_0^\infty \frac{x^4dx}{(x^2+1)^2} - \int_0^\infty \frac{x^4dx}{x^2+1}\right\}$$

$$\int_0^\infty \frac{x^2}{x^2+\alpha^2}\,dx = \left(x - \alpha\tan^{-1}\frac{x}{\alpha}\right)\Bigg|_0^\infty$$

$$\int_0^\infty \frac{x^2}{(x^2+1)^2}\,dx = \frac{-x}{2(x^2+1)}\Bigg|_0^\infty + \frac{1}{2}\cdot\frac{\pi}{2}$$

$$\int_0^\infty \frac{x^2}{(x^2+1)}\,dx = \left(x - \tan^{-1}x\right)\Bigg|_0^\infty$$

$$\int_0^\infty \frac{x^4}{x^2+\alpha^2}\,dx = \int_0^\infty x^2 dx - \alpha^2 \int_0^\infty \frac{x^2 dx}{x^2+\alpha^2}$$

$$= \int_0^\infty x^2 dx - \alpha^2 \left(x - \alpha \tan^{-1} \frac{x}{\alpha} \right) \Big|_0^\infty$$

$$\int_0^\infty \frac{x^4}{(x^2+1)^2}\,dx = \int_0^\infty \frac{x^2 dx}{(x^2+1)} - \int_0^\infty \frac{x^2 dx}{(x^2+1)^2}$$

$$= \left(x - \tan^{-1} x + \frac{x}{2(x^2+1)} \right) \Big|_0^\infty - \frac{\pi}{4}$$

$$\int_0^\infty \frac{x^4}{x^2+1}\,dx = \int_0^\infty x^2 dx - \int_0^\infty \frac{x^2}{x^2+1}\,dx$$

$$= \int_0^\infty x^2 dx - (x - \tan^{-1} x) \Big|_0^\infty$$

As a result,

$$\omega_H \int_0^\infty \frac{x^2 dx}{(x^2+\alpha^2)(x^2+1)^2} = \frac{\pi\omega_H}{4} \left(\frac{1-\alpha}{1-\alpha^2} \right)^2$$

$$\omega_H^3 \int_0^\infty \frac{x^4 dx}{(x^2+\alpha^2)(x^2+1)^2} = \frac{\pi\omega_H^3}{4} \left(\frac{1-\alpha}{1-\alpha^2} \right)^2 (1+2\alpha)$$

If B_n (noise bandwidth) is defined as $\frac{\pi f_H}{4}\left(\frac{1-\alpha}{1-\alpha^2}\right)^2$

$$\overline{i_n^2} = 4kT\left\{\frac{B_n}{R_f} + \frac{64C_1^2\gamma B_n^3}{g_m}\left(\frac{1-\alpha^2}{1-\alpha}\right)^4(1+2\alpha)\right\}$$

Let $M = \left(\frac{1-\alpha}{1-\alpha^2}\right)^2$ and $N = (1+2\alpha)$ and $\gamma = 2/3$

then,

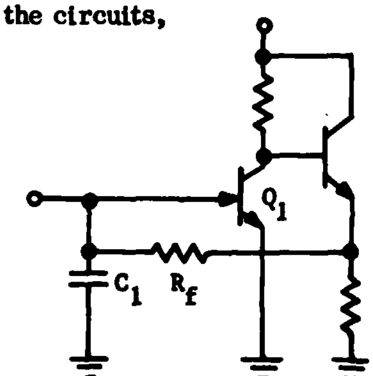
$$\overline{i_n^2} = 4kT\left(\frac{B_n}{R_f} + \frac{128C_1^2B_n^3N}{3M^2g_m}\right)$$

and,

$$\frac{i_n}{\sqrt{Hz}} = \sqrt{4kT\left(\frac{1}{R_f} + \frac{128C_1^2B_n^2N}{3M^2g_m}\right)}$$

B.2 EQUATION FOR NOISE IN A BIPOLAR TRANSISTOR FRONT END TRANSIMPEDANCE AMPLIFIER

For the circuits,



Q_1 = 2N2857
β = 100
C_1 = 5 pF

$$\overline{i_n^2} = 4kT\left\{\left(\frac{1}{R_f} + \frac{qI_e}{2kT\beta}\right)B_n + 64C_1^2B_n^3\frac{N}{M^2}\left(\frac{kT}{2qI_e} + r'_{bb}\right)\right\}$$

This was developed in the same manner as the equation for the JFET except that for a bipolar transistor,

$$\Phi(\omega) = \frac{4kT}{R_f} + \frac{2qI_e}{\beta} + \omega^2 C_1^2 \left[\frac{2(kT)^2}{qI_e} + 4kTr'_{bb} \right]$$

where I_E is the emitter current, r'_{bb} is the base resistance, and β is the current gain of the transistor.

The optimum I_E for minimum noise is

$$I_E = \frac{8C_1 kTB_n}{Mq} \sqrt{\beta N}$$

Substituting,

$$\overline{i_n^2} = 4kT \left\{ \frac{B_n}{R_f} + \frac{8C_1 B_n^2}{\sqrt{\beta}} \frac{\sqrt{N}}{M} + \frac{64C_1^2 NB_n^3 r'_{bb}}{M^2} \right\}$$

and

$$\frac{i_n}{\sqrt{Hz}} = \sqrt{4kT \left[\frac{1}{R_f} + \frac{8C_1 B_n \sqrt{N}}{\sqrt{\beta}\, M} + \frac{64C_1^2 NB_n^3 r'_{bb}}{M^2} \right]}$$

B.3 EQUATIONS FOR GIVEN CIRCUIT VALUES

For JFET full-width Manchester,

$$\alpha = 0,\ M = N = 1$$

$$i_n = \left(1.134 \times 10^{-10}\right) \sqrt{\frac{f_H}{R_f} + \left(6.58 \times 10^{-19}\right) f_H^3}$$

For Bipolar PPM Manchester,

$\alpha = .125,\ M = 0.79,\ N = 1.25$

$$i_n = \left(1.008 \times 10^{-10}\right)\sqrt{\frac{f_H}{R_f} + \left(3.5 \times 10^{-12}\right)f_H^2 + \left(6.16 \times 10^{-20}\right) f_H^3}$$

if $f_H = 8 \times 10^6$,

$i_n = 1.95 \times 10^{-9}$

PROPOSED

MILITARY STANDARD

AIRCRAFT INTERNAL TIME

DIVISION FREE ACCESS

MULTIPLEX DATA BUS

This draft Military Standard was prepared by IBM Corporation under Navy contract N00123-77-C-0747 for the Naval Ocean Systems Center (NOSC), San Diego, California.

1. SCOPE AND PURPOSE.

1.1 Scope. This standard defines requirements for a distributed, digital, free access, time division multiplexed fiber optic data bus. It defines the concepts of operation and information flow on the bus, the electrical, optical, and functional formats to be employed, and in the appendix describes certain aspects of the standard in a general sense intended to provide the user more insight into the aspects discussed.

1.2 Purpose. The purpose of this document is to establish uniform requirements for a distributed multiplex data system which will be utilized in the integration of physically distributed subsystems to operate in concert as a total system and to promote standard digital interfaces for these associated subsystems where they attach to and use the data bus. As with many standards, subtle differences will exist in its application to a given system because of the options allowed in the standard. The system designer must accommodate these differences in the design of the bus terminal hardware and software. These designer options must exist to allow the flexibility required in assembling a custom multiplex system from functionally stardard parts and to program the standard electronic functions to provide control mechanisms, traffic patterns, redundancy and fail soft concepts.

2. APPLICABLE DOCUMENTS. (TBD)

3. DEFINITIONS.

3.1 Bit. Contraction of binary digit: may be either zero or one. In information theory a binary digit is equal to one binary decision or the designation of one of two possible values or states of anything used to store or convey information.

3.2 Bit rate. The number of bits transmitted per second.

3.3 Pulse code modulation (PCM). The form of modulation in which a signal is sampled, quantized, and coded so that each element of information consists of different types or numbers of pulses and spaces.

3.4 Time division multiplexing (TDM). The transmission and exchange of information among several terminal points through one communication system with each exchange staggered in time to form a composite series of pulse trains.

3.5 Half duplex. Operation of a data transfer system between two or more terminals in either direction but not both directions simultaneously.

3.6 Word. In this document a word is a sequence of sync (2 bit times), information (17 bits), and parity (1 bit). There are five types of 20 bit words: address, command, status, priority, and data.

3.7 Message. On this data bus, a message is the transmission of a series of data words from one terminal to one or all of the other terminals on the bus.

3.8 Subsystem. The device or functional unit receiving data transfer service from the data bus.

3.9 Data bus. Whenever a data bus or bus is referred to in this document, it shall imply all of the hardware required to provide a single serial data path among all of the terminals on the bus.

3.10 Transmit/receive unit (T/RU). The hardware or module necessary to interface the data bus with the BIU specified in 3.11. T/RUs together with BIUs may exist as separate line replaceable units (LRUs), or may be contained within the elements of the subsystem.

3.11 Bus interface unit (BIU). The hardware or module necessary to interface the Transmit/Receive Unit with the subsystem. BIUs may exist with T/RUs as separate line replaceable units (LRUs) or may be contained within the elements of the subsystem.

3.12 Terminal. That LRU containing a T/RU and BIU.

3.13 Bus monitor. That terminal assigned the task of listening to bus traffic and extracting selected information to be used at a later time.

3.14 Asynchronous operation. For the purpose of this standard, asynchronous operation is the use of an independent clock source in each terminal for message transmission. Decoding may be achieved in receiving terminals using clock information derived from the message.

3.15 Bus protocol. That set of formalized sequences or exchange of words between one or more terminals on the bus in order to establish conditions for and complete the transmission of a message from one terminal to one or all of the others.

3.16 Free access. A bus protocol which allows any terminal to access and control the bus under a prioritized system where each terminal has an assigned priority different from the others. A free access bus can also function in a Command/Response or Polled Contention mode. Figure 1 shows a sample free access multiplex data bus architecture.

3.17 Redundant data bus. The use of more than one data bus to provide more than one path between the subsystems; i.e., dual redundant, tri-redundant, etc.

3.18 Broadcast. Operation of a data bus such that the transmitting terminal addresses all of the other terminals on the bus.

3.19 Bus controller. That terminal which has just accessed the bus by a procedure specified in 4.3.3.7.

3.20 Polling response frame. A pulse transmitted by a terminal in response to a priority polling procedure as specified in 4.3.3.7 and having characteristics as specified in 4.3.3.6.

4. REQUIREMENTS.

4.1 Test and operating requirements. All requirements as specified herein shall be valid over the environmental conditions in which the multiplex data bus system shall be required to operate.

4.2 Data bus operation. The multiplex data bus system in its most elemental configuration shall be as shown in Figure 1. It shall function asynchronously in a free-access mode and transmission shall occur in a half-duplex manner. Control of information transmission on the bus shall reside with the terminal which has successfully accessed the bus by means described in 4.3.3.7. The information flow on the data bus shall be comprised of word and message sequences which are, in turn, formed by five types of words (address, command, status, priority, and data) as defined in 4.3.3.5.

4.3 Characterstics.

4.3.1 Data form. Digital data may be transmitted in any desired form, provided that the chosen form shall be compatible with the message and word formats defined in this standard. Any unused bit positions in a word shall be transmitted as logic zeros.

4.3.2 Bit priority. The most significant bit shall be transmitted first with the less significant bits following in descending order of value in the data word. The number of bits required to define a quantity shall be consistent with the resolution or accuracy required. In the event that multiple precision quantities (information accuracy or resolution requiring more than 16 bits) are transmitted, the most significant bits shall be transmitted first, followed by the word(s) containing the lesser significant bits in numerical descending order. Bit packing of multiple quantities in a single data word is permitted.

4.3.3 Transmission method.

4.3.3.1 Modulation. The optical signal shall be transferred over the data bus in serial digital pulse code modulation form.

4.3.3.2 Data code. The data code at the BIU - T/RU interface shall be Manchester II bi-phase level. In the absence of data, a down level shall be transmitted. A logic one shall be transmitted as a bi-level coded signal 1/0 (i.e., an up level followed by a down level). A logic zero shall be transmitted as a bi-level coded signal 0/1 (i.e., a down level followed by an up level) (see Figure 2).

4.3.3.3 Transmission rate. The transmission rate on the bus shall be 50 megabits per second with a combined accuracy and long-term stability of ±.05 percent (i.e., 25 Kb). The short term stability (i.e., stability over 1.0 second interval) shall be at least ±.005 percent (i.e., 2.5 Kb).

4.3.3.4 Word size. The word size shall be 17 bits plus the sync waveform and the parity bit for a total of 20 bit times as shown in Figure 3.

4.3.3.5 Word formats. The word formats shall be as shown in Figure 3 for the address, command, status, priority, and data words. The first four are identified as control words.

4.3.3.5.1 Address word. An address word shall be comprised of a sync waveform, followed by a format bit, word type, a high order address, a subaddress, and a parity bit (see Figure 3).

4.3.3.5.1.1 Sync (bit 1-2). The sync waveform shall be an invalid Manchester waveform as shown in Figure 4a. The width shall be two bit times, with the waveform being an up level for the first one and one-half bit times, and then a down level for the following one-half bit time. If the next bit following the sync is a logic zero, then the last half of the sync waveform will have an apparent width of one clock period due to the Manchester encoding.

4.3.3.5.1.2 Format bit (bit 3). A one in bit position 3 identifies the word as control word.

4.3.3.5.1.3 Word type (bit 4, 5). Zeros in bit positions 4 and 5 identify the word as an address word.

4.3.3.5.1.4 High order address (bit 6-11). These six bits shall be a terminal address (up to 63). Each terminal shall be assigned one or more unique addresses. Decimal address zero (000000) shall not be assigned and is lowest priority. Decimal address 63 (111111) is the highest priority.

4.3.3.5.1.5 Subaddress (bit 12-19). These eight bits shall be used for a terminal subaddress as dictated by individual terminal requirements.

4.3.3.5.1.6 Parity (bit 20). The last bit in the word shall be used for parity over the preceding 17 bits. Odd parity shall be utilized.

4.3.3.5.2 Command word. A command word shall be comprised of a sync waveform, a format bit, followed by a zero and a one in bit positions 4 and 5 respectively, a command format, a command, and a parity bit (see Figure 3).

4.3.3.5.2.1 Sync. The sync waveform shall be as specified in 4.3.3.5.1.1.

4.3.3.5.2.2 Format bit (bit 3). As specified in 4.3.3.5.1.2.

4.3.3.5.2.3 Word type (bit 4, 5). A zero in bit position 4 and a one in bit position 5 identify the word as a command word.

4.3.3.5.2.4 Command format (bit 6, 7).

4.3.3.5.2.4.1 Command Immediate (CMD IMED). A zero in bit 6 and a zero in bit 7 identify a command immediate type command word which allows transfer of two hexidecimal digits or one 8 bit binary digit of data with the command instead of requiring a data transfer sequence.

4.3.3.5.2.4.2 Control system command (CTL SYS). A zero in bit 6 and a one in bit 7 identify a control system type command word which sets up modes, resets or clears system, etc.

4.3.3.5.2.4.3 Short block command (SB CMD). A one in bit 6 and a zero in bit 7 identify a short block command word which gives two hexidecimal or 8 binary digits of word count immediately.

4.3.3.5.2.4.4 Long block command (LB CMD). A one in bit 6 and bit 7 identifies a long block command, the definition of which (TBD) (designer's discretion).

4.3.3.5.2.5 Command (bit 8-19). These twelve bits allow up to 4096 commands or 16 commands and 8 bits of data (see Figure 3). Command codes assigned are shown in Table (TBD).

4.3.3.5.2.6 Parity (bit 20). The least significant bit in the command word shall be utilized for parity as specified in 4.3.3.5.1.6.

4.3.3.5.3 Status word. A status word shall be comprised of a sync waveform, a format bit, followed by a one and a zero in bit positions 4 and 5 respectively, fourteen status bits, and a parity bit.

4.3.3.5.3.1 Sync. The sync waveform shall be as specified in 4.3.3.5.1.1.

4.3.3.5.3.2 Format bit (bit 3). As specified in 4.3.3.5.1.2.

4.3.3.5.3.3 Word type (bit 4, 5). A one in bit 4 and a zero in bit 5 identify the word as a status word.

4.3.3.5.3.4 Status bits (bit 6-19).

4.3.3.5.3.4.1 Buffer bit (bit 6). A zero in this bit identifies the terminal as having a maximum buffer of 4 words. A one in this bit identifies the terminal as a fully buffered device.

4.3.3.5.3.4.2 Busy bit (bit 7). This bit shall be reserved for the busy bit. This bit, when used, shall indicate that the terminal is unable to accept or transmit data between the bus and the subsystem. A logic one shall indicate a busy condition, and a logic zero its absence. In the event the busy bit is set in response to a command, the terminal shall transmit its status word only.

4.3.3.5.3.4.3 End bit (bit 8). A one in this bit indicates the end of a transaction by the terminal transmitting the status word.

4.3.3.5.3.4.4 Status bits (bit 9-19). (TBD)

4.3.3.5.3.4.5 Parity bit (bit 20). The least significant bit in the status word shall be utilized for parity as specified in 4.3.3.5.1.6.

4.3.3.5.4 Priority word. A priority word shall be comprised of a sync waveform, a format bit, followed by ones in bit positions 4 and 5, a priority polling field, eight reserved bits, and a parity bit (see Figure 3).

4.3.3.5.4.1 Sync. The sync waveform shall be as specified in 4.3.3.5.1.1.

4.3.3.5.4.2 Format bit (bit 3). As specified in 4.3.3.5.1.2.

4.3.3.5.4.3 Word type (bit 4, 5). Ones in bit positions 4 and 5 identify the word as a priority word.

4.3.3.5.4.4 Priority mask (bit 6-11). These six bits are bit significant and only one bit at a time can be a logic one with the remaining five set to zero. Each bit in the priority mask identifies one of the six bits of the terminal priority number. Bit 6 of the priority mask identifies the most significant bit of the terminal's priority number and bit 11 the least significant bit. The bus access means using priority polling is specified in 4.3.3.7.

4.3.3.5.4.5 Reserved bits (bit 12-19). (TBD)

4.3.3.5.4.6 Parity bit (bit 20). The least significant bit in the priority word shall be utilized for parity as specified in 4.3.3.5.1.6.

4.3.3.5.5 Data word. A data word shall be comprised of a sync waveform, a format bit, data bits, and a parity bit (see Figure 3).

4.3.3.5.5.1 Sync. The sync waveform shall be as specified in 4.3.3.5.1.1.

4.3.3.5.5.2 Format bit (bit 3). A zero in bit position 3 identifies the word as a data word.

4.3.3.5.5.3 Data (bit 4-19). The sixteen bits following the format bit shall be utilized for data transmission as specified in 4.3.2.

4.3.3.5.5.4 Parity (bit 20). The least significant bit in the data word shall be utilized for parity as specified in 4.3.3.5.1.6.

4.3.3.6 Polling response frame. The polling response waveform shall be an invalid Manchester waveform as shown in Figure 4b. The width shall be 3 bit times, with the waveform being an up level for the first two and one-half bit times and then down for the following one-half bit time.

4.3.3.7 Data bus access. A terminal shall access the data bus by a procedure controlled by the last terminal to have the bus controller function. When this terminal completes the transaction for which it obtained control of the bus, it will transmit a priority word with the most significant priority mask bit (bit 6) set to a one and the remainder (bit 7-11) set to zero. All of the terminals (including the current bus controller) will decode the priority word and begin the data bus access sequence. If a terminal does not wish to access the bus, it will do nothing during the priority sequence. A terminal wishing to access the bus, and having a one in the most significant bit of its priority code, and after a response time as specified in 4.3.3.10, will transmit a single polling response frame as specified in 4.3.3.6. A terminal wishing to access the bus, and having a zero in the most significant bit of its priority code, will drop out of the access procedure if, at any time during a gap time as specified in 4.3.3.12, a polling response frame appears on the bus. If a polling response frame does not appear during this time, the terminal will remain in contention. After this last gap time, the current bus controller will transmit a second priority word with priority mask bit 7 set to a one and bits 6, 8-11 set to zero. Those terminals remaining in contention for the bus will repeat the access procedure just described for their second most significant priority bit. This access procedure will continue as many times as there are bits in the assigned priority number. At the end of the procedure only one terminal will remain in contention. This terminal may now access and control the bus as specified in 4.3.3.8. If no terminal gained access, the procedure is repeated until the bus has been assigned. (Note that when more than one terminal responds with a polling response frame, the pulses on the bus may overlap because of physical terminal spacing and appear to last longer than a normal polling response frame.)

4.3.3.8 Message formats. The messages transmitted on the data bus shall be in accordance with the formats in Figure 5. The allowable gap and response times shall be as stated in 4.3.3.9, 4.3.3.10, and 4.3.3.11. No message formats, other than those defined herein, shall be used on the bus. All sequences as specified herein shall begin after the data bus access procedure specified in 4.3.3.7.

4.3.3.8.1 Sequences with no data words (4.3.3.5.5).

4.3.3.8.1.1 Immediate sequence. The bus controller transmits an address word with a high order address and a sub address. The terminal addressed responds with an identical address word. The controller transmits a command with immediate data. The terminal addressed responds with status. The sequence ends.

4.3.3.8.1.2 Status polling sequence. The bus controller transmits an address word with a high order address. The terminal addressed responds with an identical address word. The bus controller transmits a command to test device. The terminal addressed responds with status. The sequence ends.

4.3.3.8.2 Sequences with data words (4.3.3.5.5).

4.3.3.8.2.1 Write sequences.

4.3.3.8.2.1.1 Write-buffered-busy. The bus controller transmits an address word. The terminal addressed responds with an identical address word. The bus controller transmits a write command. The terminal addressed responds with a status word identifying itself as a fully buffered device and with the busy bit set to one. The sequence ends.

4.3.3.8.2.1.2 Write-buffered-not busy. The bus controller transmits an address word. The terminal addressed responds with an identical address word. The bus controller transmits a write command. The terminal addressed responds with a status word identifying itself as a fully buffered device and with the busy bit set to zero. The bus controller transmits a block of data words. The terminal addressed responds with status. The sequence ends.

4.3.3.8.2.1.3 Write-unbuffered-busy. The bus controller transmits an address word. The terminal addressed responds with an identical address word. The bus controller transmits a write command. The terminal addressed responds with a status word identifying itself as an unbuffered device (4 word buffer only) and with its busy bit set to one. The bus controller starts a programmable busy bit count in order to avoid a busy hang condition on the bus and retransmits the write command. The terminal addressed responds with a status word as before with the busy bit still set to one. This sequence continues for the programmed busy bit count, and when the programmed count is reached, the bus controller sends an immediate disconnect command to the terminal address and the sequence ends. If during the sequence, and before the busy count is reached, the terminal addressed responds with its busy bit set to zero, the bus controller will transmit 4 data words and reset its busy bit counter. The receiving terminal will respond with another status word as before (busy or not busy). Until the busy counter overflows, or until the word count is reached, this sequence will continue. When the busy count or the word count is reached, the sequence will end with the receiving terminal transmitting an end status.

4.3.3.8.2.1.4 Write-unbuffered-not busy. The bus controller transmits an address word. The terminal addressed responds with an identical address word. The bus controller transmits a write command. The terminal addressed responds with a status word identifying itself as an unbuffered device (4 word buffer only) and with its busy bit set to zero. The bus controller starts a programmable busy bit counter and transmits 4 data words. The receiving terminal responds with a status word as before (busy or not busy). The procedure then continues as specified in 4.3.3.8.2.1.3.

4.3.3.8.2.2 Read sequences.

4.3.3.8.2.2.1 Read-buffered-busy. The bus controller transmits an address word. The terminal addressed responds with an identical address word. The bus controller transmits a read command and identifies itself as a fully buffered device. The terminal addressed responds with a status word with its busy bit set to one. The sequence ends.

4.3.3.8.2.2.2 Read-buffered-not busy. The bus controller transmits an address word. The terminal addressed responds with an identical address word. The bus controller transmits a read command and identifies itself as a fully buffered device. The terminal addressed responds with a block or blocks of data words (4 at a time if unbuffered) with a gap between blocks equal to or less than that specified in 4.3.3.9. The sequence ends if the gap time is exceeded. If the gap time is never exceeded, the bus controller will transmit a status word when the word count is reached. The sequence ends.

4.3.3.8.2.2.3 Read-unbuffered-busy. The bus controller transmits an address word. The terminal addressed responds with an identical address word. The bus controller transmits a read command and identifies itself as an unbuffered device (4 word buffer only). The terminal addressed responds with a status word with its busy bit set to one. The bus controller starts its busy bit counter and retransmits the read command. If the terminal addressed is still busy, the busy loop continues until the busy count is reached at which time the sequence ends.

4.3.3.8.2.2.4 Read-unbuffered-not busy. The bus controller transmits an address word. The terminal addressed responds with an identical address word. The bus controller transmits a read command and identifies itself as an unbuffered device (4 word buffer only). The terminal addressed transmits 4 words. The bus controller responds with status (busy or not busy). If this status is busy, the terminal addressed responds with busy status and starts its busy bit counter. If the bus controller is still busy, it transmits another busy status. This continues

until the busy count is reached and the sequence ends or until the bus controller is not busy and sends a not busy status to the terminal addressed. If the terminal addressed is now busy, the sequence as specified in 4.3.3.8.2.2.3 is initiated. If the terminal addressed is not busy, it transmits 4 more words. These sequences continue until the word count is reached at which time the bus controller transmits end status and the sequence ends.

4.3.3.8.3 Broadcast. The bus controller, instead of sending an address word after accessing the bus, transmits a command word with up to 14 bits of data. All other terminals, seeing the command word instead of an address word, recognize a broadcast mode and accept the immediate data. The sequence ends.

4.3.3.9 Intermessage or interword gap. A terminal receiving messages on the bus shall allow a programmable gap time between messages or words as shown in Figure 6. The time is measured from the mid bit time of the last bit of the preceding message or word to the negative going transition of the next word sync. (Note: This programmable time must be the same for all terminals on the bus.)

4.3.3.10 Response time. A terminal shall respond, in accordance with 4.3.3.8, to another terminal within a time period which is programmable. The time is measured from the mid bit time of the last bit of the last received word, as specified in 4.3.3.8, to the negative going transition of the first responding word sync. (Note: This programmable time must be the same for all terminals on the bus.)

4.3.3.11 Minimum no-response time-out. The minimum time that a terminal shall wait before considering that a response as specified in 4.3.3.10 has not occurred shall be as specified in 4.3.3.12. The time is measured from the mid bit time of the last bit of the last word to the negative going transition of the first expected response word sync.

4.3.3.12 Bus access gap time. The minimum bus access gap time shall be programmable and depend on the length of data bus transmission media between the most widely separated terminals. It shall be determined by the following formula:

$$t_{ag} \geq \frac{d}{\upsilon_p} + t_r + t_s$$

where: t_{ag} - bus access gap time

d - length of longest transmission media

υ_p - velocity of signal propagation in media

t_r - response time (4.3.3.10)

t_s - polling response frame time (60 nanoseconds)

The bus designer may add any margin he wishes, but the greater the margin the less efficient the bus access procedure will be. This bus minimum access gap time is the time it will take for the polling response frame to reach and be detected by a terminal polling for priority if the responding terminal and the polling terminal are the two most widely separated terminals on the bus. (Note: The programmable bus access time shall be the same for all terminals on the bus.)

4.4 Terminal operation. Terminals shall have common operating capabilities as specified in the following paragraphs.

4.4.1 Word validation. The terminal shall insure that each word conforms to the following minimum criteria:

a. The word beings with a valid sync.

b. The bits are in a valid Manchester II code.

c. The information field has 17 bits plus parity.

d. The word parity is odd.

4.4.2 Transmission continuity. A data block transmission may be non-contiguous as specified in 4.3.3.9. If the intermessage or interword gap is exceeded before the message word count is reached, the receiving terminal will transmit an end status with the message error bit set.

4.4.3 Unbuffered busy loop timeout. When a write/read-unbuffered-busy or not busy sequence is occurring and the busy count is reached before the word count, the receiving terminal will transmit an end status with the message error bit set.

4.4.4 Priority poll timeout. The terminal functioning as the bus monitor will have a timer which resets after each transmission of a priority polling word. This timer will be programmable, and if it times out, it identifies the fact that a priority polling frame has not occurred for some preset maximum time.

4.4.5 Invalid commands. A terminal shall respond to a command which does not fulfill the criteria of 4.4.1 by transmitting a status word with either the message error or parity bit set to one. The terminal sending the command will have a programmable counter which will allow N command retries ($N \geq 0$). If the counter overflows, the terminal will begin the priority sequence as specified in 4.3.3.7. All terminals which can initiate sequences shall have this capability.

4.4.6 Invalid address. If a terminal transmits an invalid address, after a response time as specified in 4.3.3.10, the terminal will retry the sequence again N times as determined by a programmable counter (N 0). If the counter overflows, the terminal will begin the priority sequence as specified in 4.3.3.8.

4.4.7 Invalid data. If a terminal receives invalid data, it will respond with a message error status. For unbuffered transmission (4 words), the sequence may be retried N times ($N \geq 0$).

4.4.8 Invalid status. When a terminal receives an invalid status word, the sequence ends as usual. Action to be taken is at the discretion of the system designer.

4.4.9 Invalid priority word. A terminal who perceives a priority polling word to be invalid (except the bus controller) shall drop out of the bus access procedure. The bus controller will continue the sequence until completed.

4.4.10 Invalid sequence. All sequences not specified in 4.3.3.8 are illegal. Occurrence of an illegal sequence shall invoke action as specified in (TBD).

4.4.11 No bus access. If, after the procedure specified in 4.3.3.7, no terminal accessed the bus, the terminal performing the procedure will repeat the procedure until some other terminal has accessed the bus. This condition will be recognized by having priority (000000) unassigned.

4.4.12 Illegal commands. An illegal command is a valid command as specified in 4.4.1, where the command, word count, or subaddress, etc has not been implemented in the receiving terminal. It is the responsibility of the terminal transmitting the command to assure that no illegal commands are sent out. The receiving terminal shall respond with a status word with the illegal command bit set.

4.4.13 Bus monitor operation. A terminal operating as a monitor shall listen to bus traffic and extract selected information. While operating as a monitor, the terminal shall only respond to messages containing its own unique address if one is assigned. All information obtained while acting as a monitor shall be strictly used for off-line applications or to provide any back-up terminals sufficient information to take over a failed terminal's function. The terminal operating as monitor is responsible for recovering bus operation in the event of a priority poll timeout.

4.5 Hardware characteristics.

4.5.1 Bus interface unit (BIU) characteristics.

4.5.1.1 BIU - T/RU interface. This interface shall consist of two differential pairs of electrical signals; a transmit pair and a receive pair (see Figure 7).

4.5.1.1.1 BIU transmit signal levels. The BIU output up and down voltage levels shall be between -1.1 and -1.6 to -0.9 and -1.8 volts, peak-to-peak, line-to-line when measured at point A in Figure 7.

4.5.1.1.2 BIU transmit signal waveform. The BIU transmit signal waveform, when measured at point A in Figure 7, shall have mid transition (center of peak-to-peak levels) crossing deviations which are equal to, or less than, ±1.0 nanoseconds from the ideal crossing point, measured with respect to the previous mid transition crossing point (i.e., 10 ± 1.0 nanoseconds, 20 ± 1.0 nanoseconds, 30 ± 1.0 nanoseconds, etc). The rise and fall time of this waveform shall be less than or equal to 1.5 nanoseconds when measured from levels of 10 to 90 percent of full waveform peak-to-peak, line-to-line voltage as shown in Figure 8. Any distortion of the waveform including overshoot and ringing shall remain within the up or down level limits as specified in 4.5.1.1.1 and as measured at point A in Figure 7.

4.5.1.1.3 BIU transmit signal noise. Any noise on the transmit signal differential output when the BIU is receiving or has power removed shall remain within the up or down level limits as specified in 4.5.1.1.1 and as measured at point A, Figure 7.

4.5.1.1.4 BIU transmit signal polarity. In the absence of a signal on the transmit signal output pair, the "TRUE" terminal of the differential output shall be at a down level (more negative of the two levels). When a word is transmitted, the "TRUE" terminal shall go to the up level (more positive of the two levels) at the beginning of the sync frame. These are the up and down levels as specified in 4.3.3.2.

4.5.1.1.5 BIU receive signal compatibility. The BIU shall be capable of receiving and operating with the incoming signals specified herein. It shall accept a waveform whose mid transition crossing deviation with respect to the previous mid transition crossing is ±1.5 nanoseconds (i.e., 10 ± 1.5 nanoseconds, 20 ± 1.5 nanoseconds, 30 ± 1.5 nanoseconds, etc). The BIU shall respond to an input signal whose peak-to-peak, line-to-line amplitude is within the range specified by 4.5.1.1.1. The BIU shall not respond to an input signal whose peak-to-peak, line-to-line amplitude is less than the limits specified in 4.5.1.1.1. These voltages are measured at point A, (see Figure 7).

4.5.1.1.6 Input impedance. The magnitude of the BIU receive pair line input impedance shall be 75 ± 25 ohms within the frequency range of 10 to 200 MHz. This impedance is measured from each of the pair to ground at point A in Figure 7.

4.5.1.2 BIU - subsystem interface. (TBD)

4.5.2 Transmit/receive unit (T/RU) characteristics.

4.5.2.1 T/RU - BIU interface. This interface shall be as specified in 4.5.1.1.

4.5.2.1.1 T/RU transmit signal levels. The T/RU output voltage levels shall be as specified in 4.5.1.1.1.

4.5.2.1.2 T/RU transmit signal waveform. The T/RU transmit signal waveform shall be as specified in 4.5.1.1.2.

4.5.2.1.3 T/RU transmit signal noise. The T/RU transmit signal noise shall be as specified in 4.5.1.1.3.

4.5.2.1.4 T/RU transmit signal polarity. The T/RU transmit signal polarity shall be as specified in 4.5.1.1.4.

4.5.2.1.5 T/RU receive signal compatibility. The T/RU shall be capable of receiving and operating with a signal as specified in 4.5.1.1.5.

4.5.2.1.6 Input impedance. The magnitude of the T/RU receive pair line input impedance shall be as specified in 4.5.1.1.6.

4.5.2.2 T/RU - data bus interface. The data bus covered by this standard shall be implemented as specified in 4.5.3. In accordance with that paragraph, a Transmit/Receive Unit (T/RU) shall have two optical ports; a transmit or output port, and a receive or input port.

4.5.2.2.1 Output port. The output port shall transmit an optical equivalent of the signal as specified in 4.5.2.1.5.

4.5.2.2.1.1 Output optical power. (TBD)

4.5.2.2.1.2 Output waveform. (TBD)

4.5.2.2.1.3 Output spectral characteristics. (TBD)

4.5.2.2.1.4 Output noise. (TBD)

4.5.2.2.1.5 Output connector. (TBD)

4.5.2.2.1.6 T/RU delay. (TBD)

4.5.2.2.2 Input port. The input port shall operate with a signal equivalent to that specified in 4.5.2.2.1 but attenuated a maximum of 30 dB and/or a minimum of 6 dB (i.e., a dynamic range of 24 dB).

4.5.2.2.2.1 Input optical power. (TBD)

4.5.2.2.2.2 Input waveform. (TBD)

4.5.2.2.2.3 Input noise. (TBD)

4.5.2.2.2.4 Input connector. As specified in 4.5.2.2.1.5.

4.5.2.2.2.5 T/RU delay. (TBD)

4.5.3 Data bus characteristics.

4.5.3.1 Cable. The fiber optic cable used for the bus shall be a suitably strengthened and jacketed fiber whose silica core is 200 ± 5 micrometers in diameter and whose cladding diameter has a ±5% tolerance.

4.5.3.2 Cable bandwidth.

4.5.3.2.1 Modal dispersion. The modal dispersion of the fiber shall be equal to or less than 10 nanoseconds per kilometer with a source wavelength from 800 to 900 nanometers.

4.5.3.2.2 Material dispersion. The material dispersion of the fiber shall be equal to or less than 5 nanoseconds per kilometer with a source center wavelength from 800 to 900 nanometers and a one-half power bandwidth of 60 nanometers.

4.5.3.2.3 Combined bandwidth. The combined effect of 4.5.3.1.1, and 4.5.3.1.2, and the source characteristics shall be a fiber whose bandwidth-distance characteristic is 25 meghertz-kilometers.

4.5.3.3 Cable attenuation. The cable attenuation shall be equal to or less than 10 dB per kilometer where the dB measured is relative to the power loss (i.e., $dB = 10 \log \frac{P_{in}}{P_{out}}$).

4.5.3.4 Cable termination. The cable ends shall be suitably terminated with an environmentally qualified connector whose axial alignment, concentricity, and angular displacement shall be (TBD).

4.5.3.5 Bus architecture. Figure 9 shows the bus architecture and terminal to bus coupling means. The optical transmissive star coupler is analogous to the "bus." The optical fiber cables between each of the terminals and the coupler are the means by which a terminal couples to the "bus" and are analogous to "stubs" on a wire bus. The transmissive star coupler may be active or passive, but has the characteristic that

whatever optical signal comes into the coupler is passed on or retransmitted to all terminals on the bus including the transmitting terminal. A transmissive star coupler requires two ports at the terminal; an input or receiving port, and an output or transmitting port. An N port transmissive star coupler will have 2N physical ports. The coupler may be located anywhere relative to the locations of the terminals on the bus.

4.5.3.5.1 <u>Passive transmissive star coupler</u>. Access couplers exhibit three optical power reduction mechanisms. The first is power division. Unlike data bus systems which utilize a wire interconnection, a fiber optic system divides signal power among all terminals having access to the bus. This division of power is accomplished by one or more access couplers. Thus, for an N port transmissive star coupler in which input power P_{in} is evenly distributed, the power available from any port due to this mechanism will equal P_{in}/N. This division loss, L_D, can be expressed as:

$$L_D = 10 \log \frac{1}{N} \text{ dB (optical)}$$

The second reduction mechanism results from the inefficiency of the coupler itself and is called excess loss, L_E. Excess loss can be described by:

$$L_E = 10 \log \left[1 - \frac{\sum_{i=1}^{N} P_{out_i}}{P_{in}} \right] \text{ dB (optical)}$$

where P_{in} is optical power into the coupler and P_{out_i} is the power out of the ith output port. These losses are attributable primarily to optomechanical factors, such as packing fraction, numerical aperture mismatch, scattering, and assembly tolerance losses.

Finally, power is reduced by insertion losses, L_I, at the cable-to-coupler and coupler-to-cable interfaces. These losses may be due to either connectors or splices, depending on the system assembly technique adopted, but each interface will have a characteristic loss, L_I.

Total coupler insertion loss from an input port to an output port, L_T, may be expressed by:

$$L_T = L_D + L_E + 2L_I \text{ dB (optical)}$$

Note that L_E and L_I will have statistical variations from port to port, depending on coupler manufacturing quality, which must be considered in bus system power budgeting.

4.5.3.5.2 Active transmissive star coupler. (TBD)

4.5.3.6 Comparison with wire bus networks. An optical data bus network is more analogous to a constant-current electrical network than to a constant-voltage network, such as the wire bus network described in MIL-STD-1553. Thus, the effective distribution of optical power from a bus terminal to n other terminals on the bus calls for the consideration of bus network configurations considerably different from the tee-tap daisy chain method of MIL-STD-1553. Worst-case optical losses for line replaceable unit (LRU) connectors, in-line bulkhead connectors, access couplers, cable, etc, must be combined with available optical power and receiver unit sensitivity to derive optical power budget and optical signal range (dynamic range) figures for any proposed bus configuration. The power budget must also be computed taking into account applications factors such as temperature, altitude, humidity, statistical component variations, age, and nuclear radiation over the ranges detailed in the applicable system and item specifications. Reliability, installation, maintainability, and related life cycle cost factors must also be considered, especially with regard to use of active (repeater) access coupling devices.

4.5.4 Data bus bit error rate. The bit error rate of a terminal to terminal transaction shall be equal to, or less than, 1 error in 10^9 bits. This measurement shall include errors caused by the BIU, T/RU, and data bus hardware when operating in any combination of conditions as specified in 4.5.1 and 4.5.2.

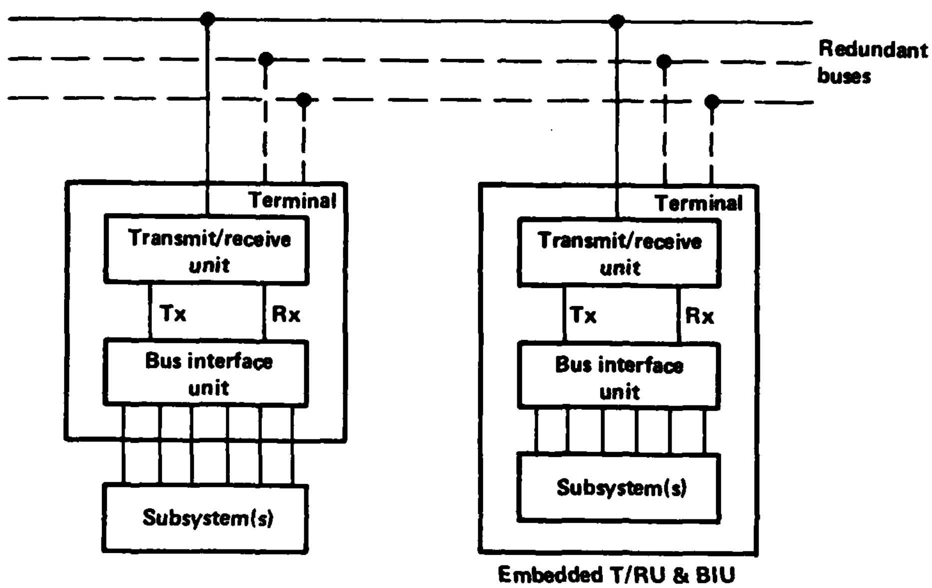


Figure 1. Sample Free-Access Multiplex Bus Architecture

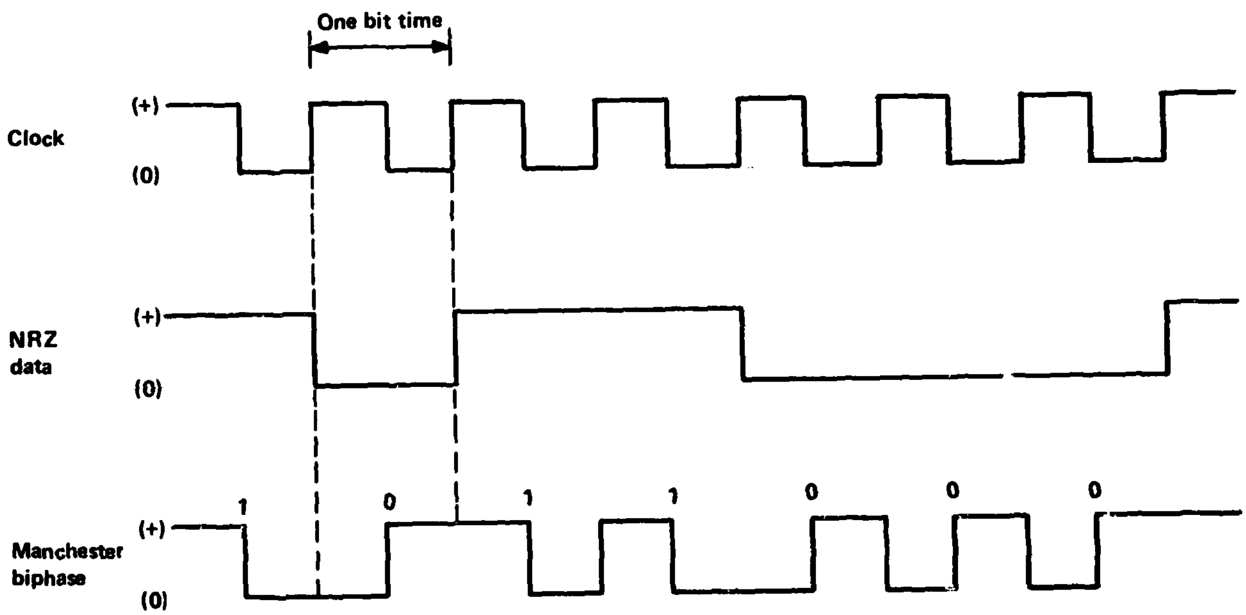


Figure 2. Data Encoding

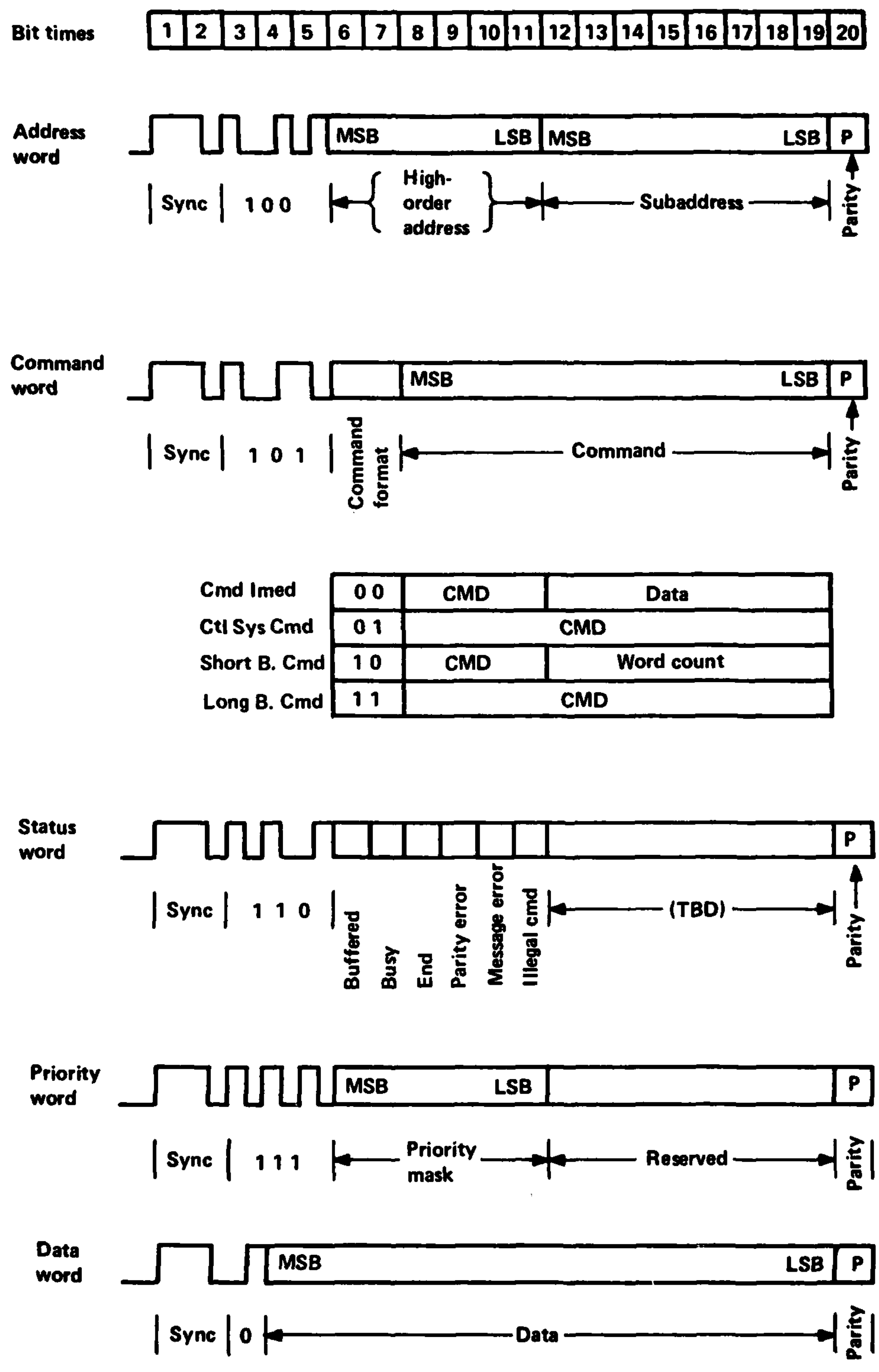


Figure 3. Word Formats

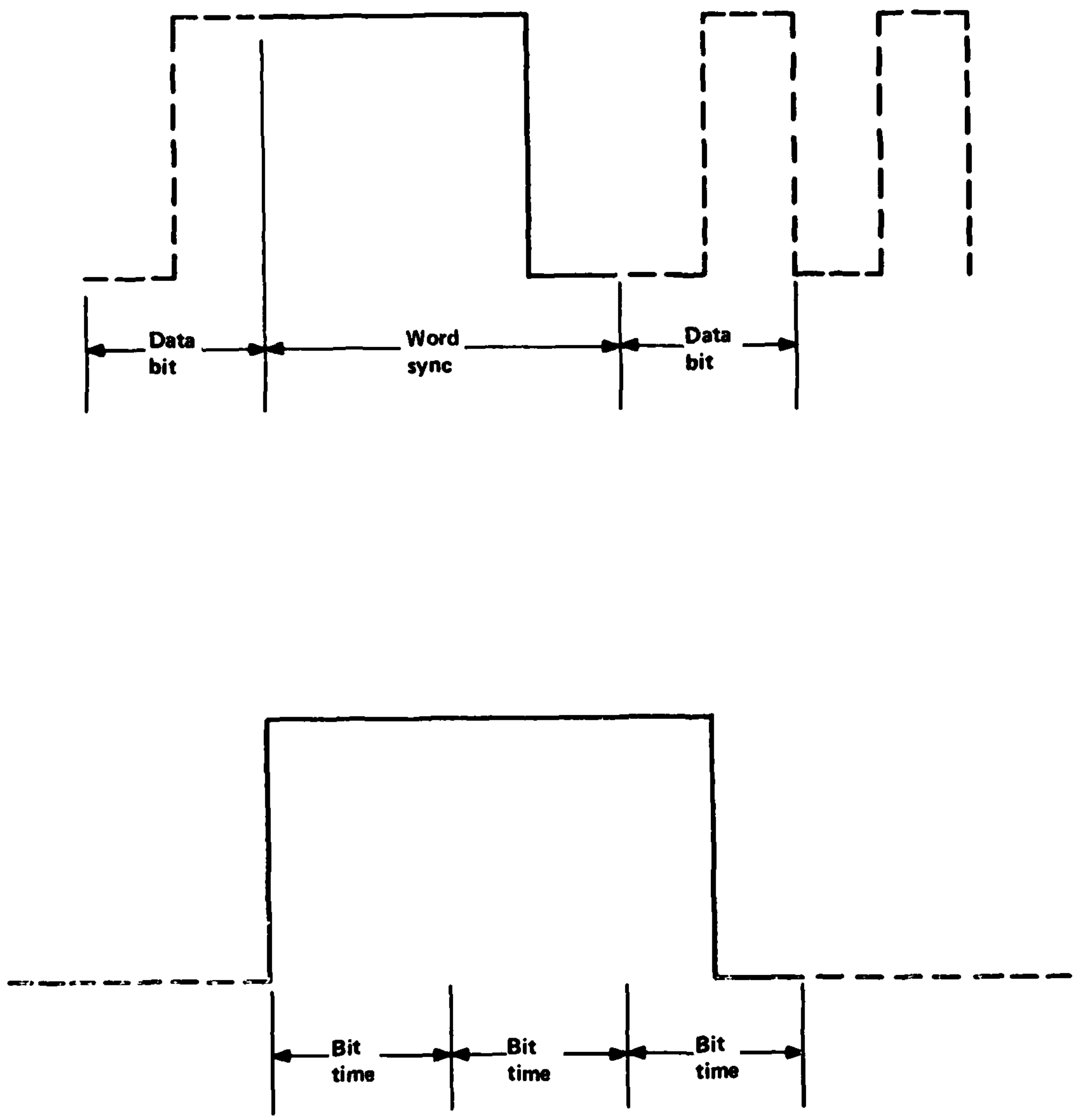


Figure 4. Word Sync and Polling Response

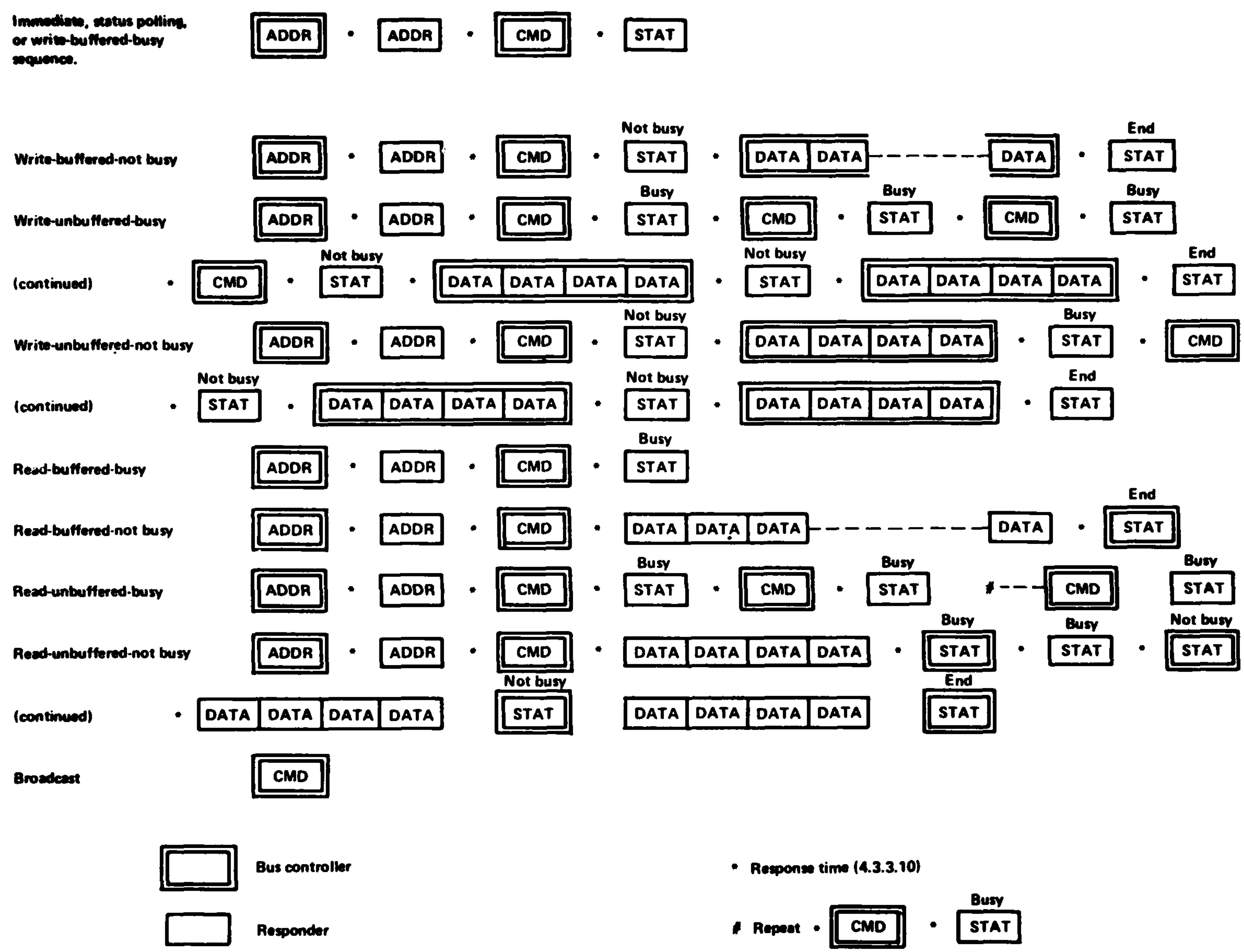


Figure 5. Message Sequences

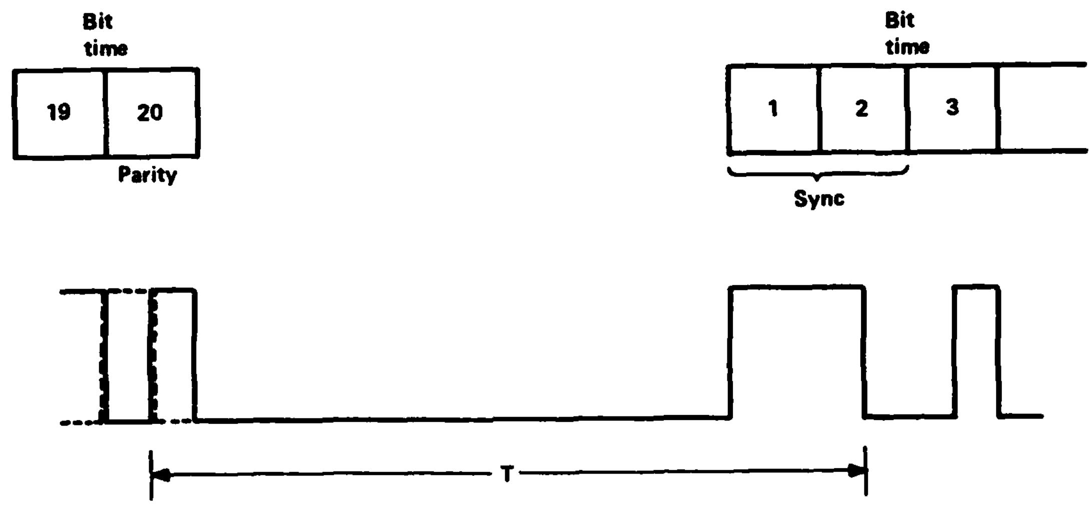


Figure 6. Gap and Response Time

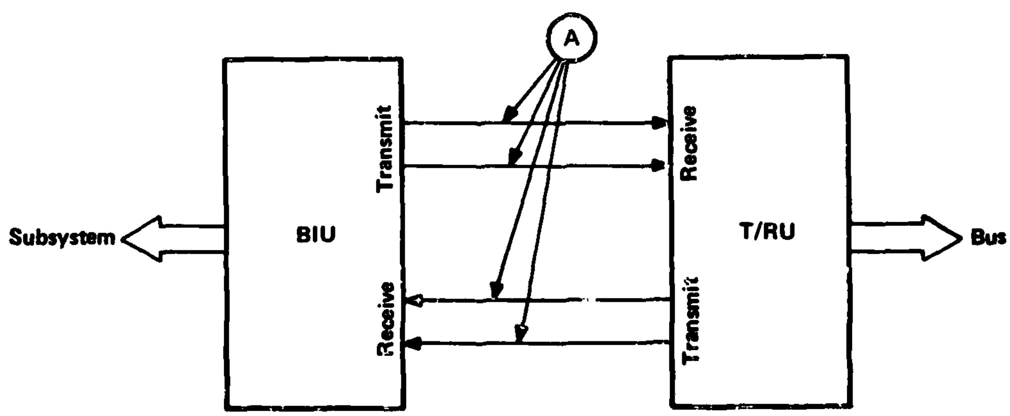


Figure 7. BIU-T/RU Interface

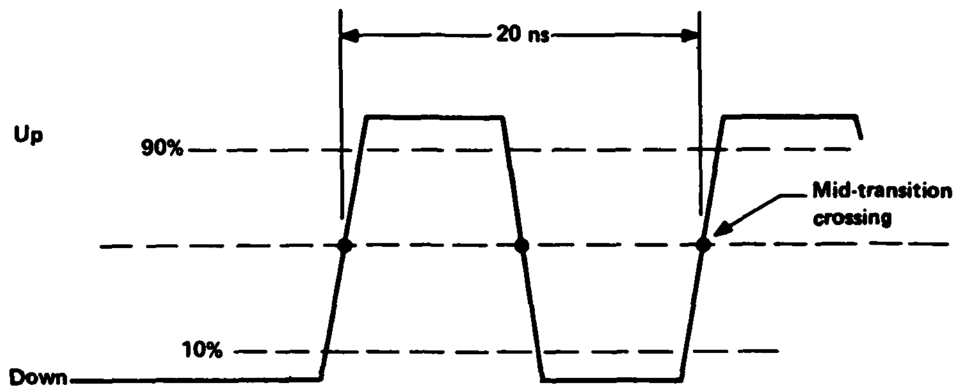


Figure 8. Interface Waveform (BIU-T/RU)

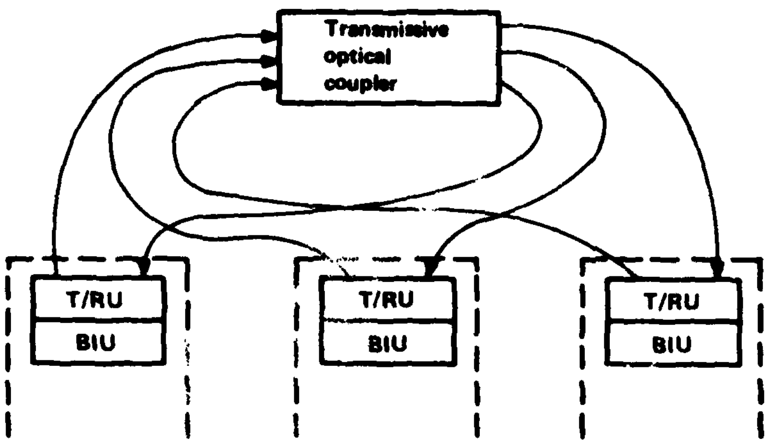


Figure 9. Fiber-Optic Bus Architecture

ANNEX

A1. Bus efficiency. Any communications link or bus may have inefficiencies because of the protocol overhead; especially a single line serial bus such as the bus in this standard. The inefficiency is greatest for messages with small word counts. In a high speed bus, another factor begins to add to the inefficiency and that is the actual propagation time of the signal between terminals. For example, a 20 bit word at 50 Mbits takes 400 nanoseconds to be serialized and takes 500 nanoseconds to travel between terminals separated by 100 meters of fiber optics. If one defines real information as only the 16 bits of information in a data word, then the formula below describes the bus efficiency for a fully buffered and an unbuffered data transfer between two terminals 50 meters apart on a bus where the worst case terminal separation can be 100 meters. Figure A1 is a plot of these efficiencies in terms of real information rate for a 16 port and 64 port bus. Light travels approximately 2×10^8 meters per second in a fiber.

Bus Real Information Rate (R)

$$R = \frac{\overbrace{16t_bM \times \frac{1}{t_b}}^{\text{Real Data}}}{\underbrace{P(\underbrace{20t_b}_{\text{Priority Word}} + \underbrace{2t_d}_{\text{Round Trip Time (100 meters)}} + t_r + \underbrace{3t_b}_{\text{Polling Response Pulse}})}_{\text{Bus Access Procedure}} + \underbrace{4(\underbrace{20t_b}_{\text{Control Word}} + \underbrace{\frac{t_d}{2}}_{\text{Bus Avg. Trip Time (50 meters)}} + t_r)}_{\text{Handshakes}} + \underbrace{\underbrace{20t_bM}_{\text{Data Words}}}_{\text{Data Words}} + \underbrace{K(\underbrace{20t_b}_{\text{Status Word}} + \underbrace{t_d}_{\text{Two Avg. Trips}} + \underbrace{2t_r}_{\text{Two Responses}})}_{\text{Statu}}}$$

M - Message Length in Words
P - Priority Polling Frames
t_b - Bit Time
t_d - Worst Case Time Between Terminals
t_r - Terminal Response Time

$K = 1$ for buffered

$K = \frac{M}{4}$ for unbuffered

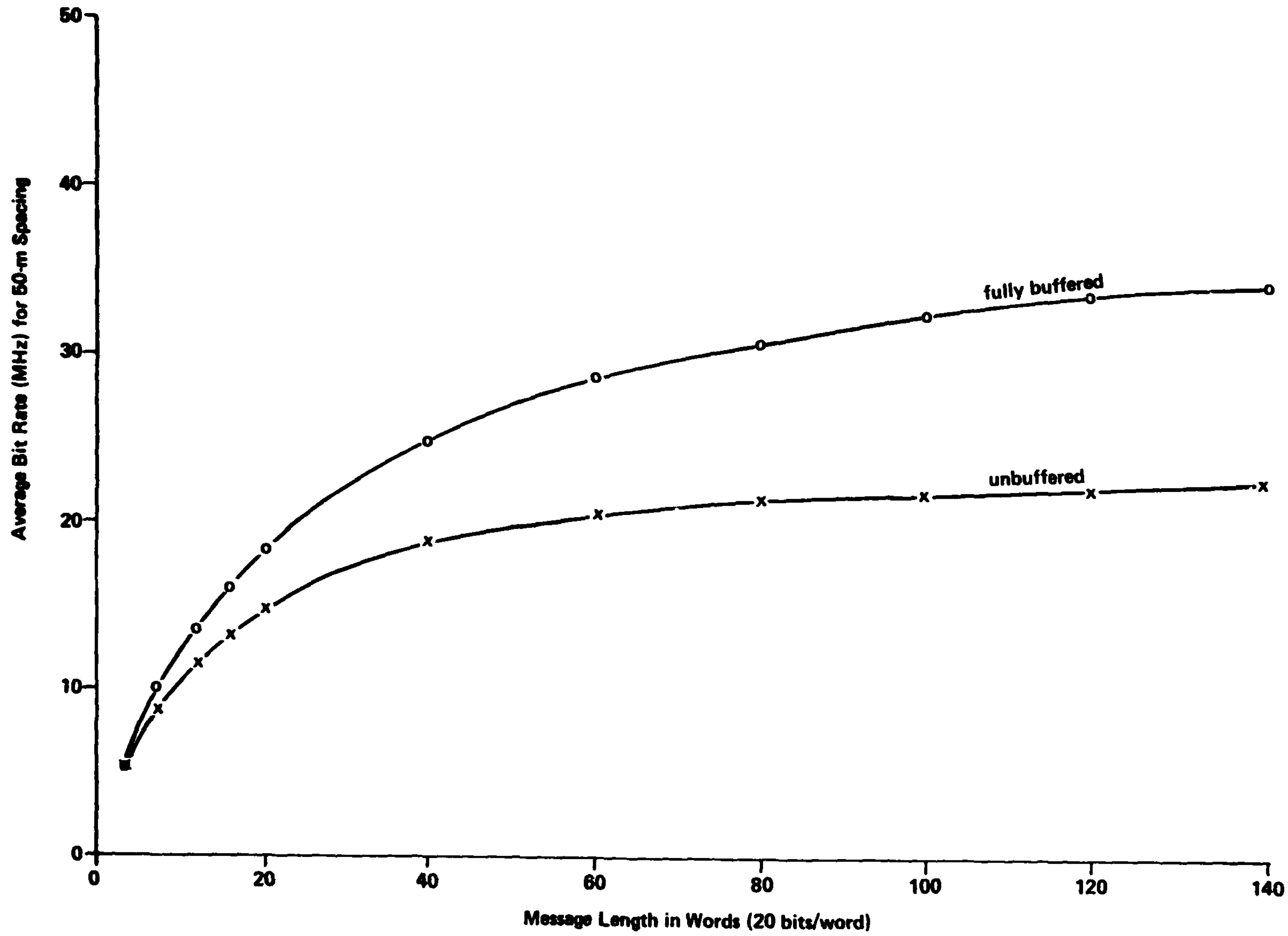


Figure A1. Real Information Rate for 50 Mb/s Data Bus With Overlay Priority Resolution

MED

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